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INVESTIGATION OF THE USE OF FREQUENCY-DIVISION MULTIPLE ACCESS FOR APPLICATION WITH A MIX OF USER TERMINALS

Volume Three — Multiple-Access Techniques

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Prepared for:

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ABSTRACT

The three major multiple-access techniques are compared for a mix of user types--i.e., differing data rates and capacity quotients. The comparison is made with respect to eight performance criteria that are particularly important for the military environment. A general comparison indicates that frequency-division multiple access (FDMA) performs better than originally believed. A specific test case of 10 accesses with a variety of data rates and capacity quotients is selected for detailed evaluation. The quantitative comparison is based on satellite throughput. The performance of the FDMA system is optimized and evaluated using the computer program SYSCON developed under the other tasks of this study. FDMA is found to offer ver[illegible]y as much satellite throughput as the other two multiple-access techniques. Furthermore, FDMA offers the advantages of simplicity, low cost, and compatibility with existing equipment. Based on this example, it is concluded that FDMA can offer substantial advantages as compared to the other techniques. Further, more detailed studies are recommended to fully delineate the region of FDMA superiority.

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*Recommended sections for initial reading.

III MULTIPLE-ACCESS TECHNIQUES AND CHARACTERISTICS (Continued)

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*Recommended sections for initial reading.

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GLOSSARY

ACI	Adjacent-channel interference
AJ	Anti-jam
ATDMA	Asynchronous time-division multiple access
BPSK	Binary phase-shift keying
BU	Bandwidth utilization
CDMA	Code-division multiple access
C/KT	Ratio of carrier power to noise spectral density
ERP	Effective radiated power
FDMA	Frequency-division multiple access
FHMA	Frequency-hopping multiple access
G/T	Ratio of receiver antenna gain to system noise temperature
IM	Intermodulation cross products
Mbps	Megabits per second
NOMAC	Noise modulation and correlation
PAMA	Pulse-address multiple access
PCMA	Phase-chipping multiple access
PN	Pseudo-noise
PSK	Phase-shift keying
QPSK	Quaternary phase-shift keying
RADA	Random-access discrete address
RF	Radio frequency
S/I	Ratio of signal to interference power
SPADE	COMSAT's single-channel-per-carrier FDMA system
SS	Spread spectrum
SSMA	Spread-spectrum multiple access
SSMAC	Spread-spectrum multiple access coherent

SSMAN	Spread-spectrum multiple access noncoherent
SNR	Signal-to-noise ratio
TDMA	Time-division multiple access
TP	Throughput

I INTRODUCTION

This report presents the results of the multiple-access-system comparison task of a larger study devoted to precise evaluation of the capabilities of frequency-division multiple access (FDMA). This evaluation is necessary because previous analyses of FDMA were overly pessi mistic about its capabilities.

Previous analyses of FDMA assumed analog FM modulation. As a result, when it was determined that hard-limiting satellite transponders produced signal-to-interference (S/I) ratios of 9 dB (below the threshold of conventional FM discriminators), it was concluded that FDMA was an unsatisfactory technique. However, nonuniform frequency plans can avoid cross products and increase the S/I ratio. Using only 50 percent of the available bandwidth makes possible the avoidance of one-half the cross product power and improvement of the S/I ratio by 3 dB. The use of quadriphase modulation (QPSK) rather than biphase modulation (BPSK) makes it possible to obtain (1) the same number of frequency channels in a given bandwidth, and (2) an S/I ratio of 12 dB rather than 9 dB.

In addition, it is possible to back off the input power somewhat, reducing the power in the desired signal components slightly, but significantly decreasing the power in the intermodulation cross products. Thus, depending on the thermal noise levels, it may be possible to obtain several more decibels of improvement in the S/I ratio.

Since modern signaling techniques are digital (for a variety of good reasons), we are concerned with the error-rate characteristic instead of discriminator thresholds. For BPSK or QPSK, a detector signal-to-noise spectral-density ratio (E/N_o) of 10.6 dB yields an error rate

of 10^{-6}. Even for E/N_o's as low as 6.8 dB, the error rate is less than 10^{-3}. While this value may be higher than desired, it can be readily reduced to the desired value by the use of convolutional coding and sequential decoding. Such coding systems are being used more and more in satellite communication. Thus, one concludes that it is quite reasonable to expect good performance with a QPSK/FDMA system operating in an optimized (with respect to frequency and power) manner.

Based on these observations, a study effort was undertaken to evaluate precisely the performance of an optimized QPSK/FDMA system for the realistic case of a mix of user data rates, effective radiated powers (ERPs), and receiver senitivities. This report compares the relative desirability of the three major multiple-access categories for this environment.

Section II develops the background of this work. The assumptions, the objectives, the performance criteria, and the concept of link unbalance are presented. The required technical expressions and feasibility assessments for each of the multiple-access techniques are developed in Section III. Section IV presents the comparative performance of FDMA, SSMA (spread-spectrum multiple-access), and TDMA (time-division multiple-access) systems for a mix of 10 accesses. Conclusions are given in Section V.

Due to the complexity of the multiple-access problem, this report is necessarily lengthly. A general understanding of the major points can be gained by reading Sections I, II, III-D, III-E, IV, and V, which constitute a relatively small fraction of the total report.

II BACKGROUND

A. Assumptions

This comparison of multiple-access systems is based on a set of assumptions. The first is that there will be a mix of user terminal types--i.e., effective radiated power (ERP) and antenna gain-to-noise-temperature ratio (G/T)--and data-rate requirements. Second, the application is to strategic-type communication where dedicated links are established to handle trunk traffic. These links are quite stable and require reconfiguration no more frequently than every several hours.* For very rapidly changing user population and rate requirements--e.g., data terminals and voice net traffic--it may be desirable to consider such multiple-access techniques as asynchronous time-division multiple access (ATDMA), that are not considered in this report. One form of ATDMA that appears particularly attractive for this situation is the packet switching concept employed in the ARPANET and within the ALOHA system.

Third, it is assumed that a saturating, frequency-translating, wideband transponder is used. Advances in the satellite technology that develops processing transponders or multichannel transponders will affect the analyses and conclusions of this report. In particular, processing transponders will enhance the capabilities of SSMA, whereas multichannel transponders may improve the characteristics of FDMA with respect to power control.

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*Some forms of tactical trunk communication also meet this description.

B. Objective

Based on the results of Task 3, a technical assessment was to be made of PSK/FDMA, relative to known multiple-access techniques, which include at least those multiple-access techniques based on time division, frequency division, and code division (spread spectrum). The assessment was to address those areas for which PSK/FDMA is particularly suited in terms of overall communication-system performance and operations responsiveness.

C. Performance Criteria

Selection of the best multiple-access technique is complicated by the fact that many performance criteria exist. Thus, it is necessary to select and weight each of the significant criteria in choosing the best multiple-access technique. The following are the proposed criteria, listed in order of significance, for the work to be performed under Task 4:

- Total throughput[†] of satellite transponder [measured in megabits per second (Mbps)]
- Tolerance to inequities between receiver G/T and data rate
- Flexibility--e.g., ability to accommodate nonstandard data rates
- Complexity of operational doctrine and procedures
- Anti-jam (AJ) capability
- Compatibility with existing equipment
- Equipment cost

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[*] The required error rate will be considered to be a specification that the proposed system must meet.

[†] Throughput is defined as the total bit rate through the transponder--i.e., the sum of the rates of each access.

- Robustness with respect to user errors and environment imperfections.

In the process of performing analyses and tailoring system parameters to user requirements, it is convenient to use two other criteria: energy efficiency and bandwidth utilization. Since these criteria are not useful for end-item comparisons, they are not in the above list. Nevertheless they are of sufficient working interest to be described here.

Energy efficiency is a measure of the energy required to detect a single bit with a prescribed probability of error--e.g., 10^{-6}. Binary phase-shift keying (PSK) is more efficient than octal phaseshift keying or binary on-off keying because it requires less energy to achieve the same probability of error.

Bandwidth utilization is defined as the data rate divided by the required radio frequency (RF) bandwidth. Consider the example of a single binary PSK access. It will be assumed that the required RF bandwidth must be the first-null-to-first-null bandwidth. It is acknowledged that this bandwidth restriction will result in some loss in signal detectability. However, any finite bandwidth will result in a loss; the required RF bandwidth selected is convenient to remember. For the binary PSK example, the bandwidth utilization is 0.5. For a quaternary PSK signal, the bandwidth utilization is 1.0. This concept of bandwidth utilization is extended to the multiple-access case in the obvious fashion of using total data rate and total RF bandwidth.

These two criteria can be used in the comparison of multiple-access techniques at intermediate levels. Final comparisons for test cases will use the listed system performance criteria.

D. Link Imbalances

Two types of imbalance can occur in satellite communication systems. Consider the simple example of two accesses.

First, a power imbalance can occur. This imbalance is desirable if the two transmissions with different data rates are intended for terminals with the same G/T. The higher-rate transmission clearly requires more power than the lower-rate transmission to achieve the same error rate.

Second, a power-spectral-density imbalance can occur. This imbalance is desirable if two transmissions at the same rate are intended for terminals with different G/T. The transmission intended for the lower G/T terminal requires the greater spectral density. Figure 1 illustrates spectral diagrams for these two cases.

It should be noted that both types of imbalance can occur simultaneously for a single transmission--e.g., a very-high-rate transmission intended for a very low G/T. Power-spectral-density imbalance is created by differences in receive G/T figure of merit. Power imbalance is created by differences in the (data rate)/(G/T) ratio.

Power imbalance has a significant effect on the following factors: (1) power sharing, (2) suppression, and (3) magnitude of cross products. Spectral-density imbalance has effects on the following factors: (1) density of the interference falling within the receiver passband, (2) level of spectrum sidelobes, and (3) spectral density of cross products. Thus, it is desirable to keep accurate accounting of both power and power-spectral-density imbalances in satellite links.

The case of link imbalance is of particular importance for military satellite communication systems. Commercial systems tend to encounter a much greater user uniformity. The diversity of G/T's and data rates for military communications places a much greater strain on the multiple-access

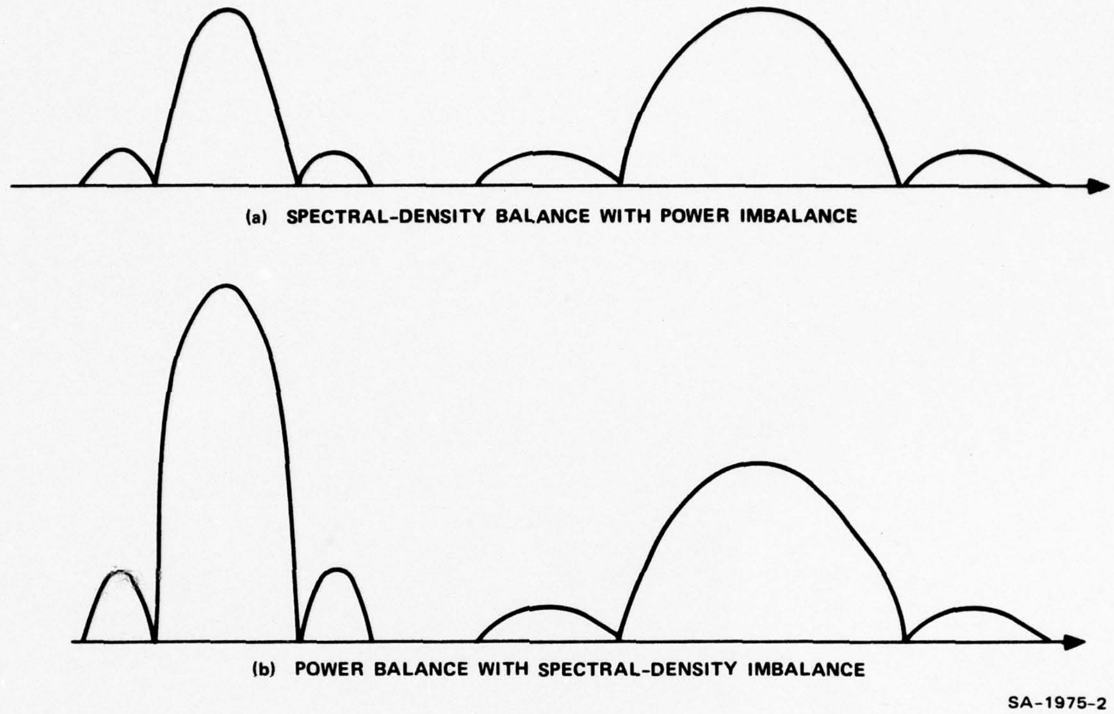

FIGURE 1 SPECTRAL DIAGRAMS

techniques. It is this variety that is responsible for the need for a more detailed multiple-access-technique comparison.

III MULTIPLE-ACCESS TECHNIQUES AND CHARACTERISTICS

This section develops the required analyses to properly compare the three multiple-access categories for the case of a mix of user types--i.e., a variety of capacity quotients and data rates. The analyses presented are restricted to items related to user differences. There is no intent to completely document all aspects of multiple-access systems. Greater breadth of coverage is available in works by other researchers.[1*]

A. Code-Division Multiple Access

In this section, code-division multiple-access (CDMA) techniques are categorized by type, and concentration is focused on the most common. Multiple-access performance is evaluated, and attention is paid to power-control and data-rate limitations. AJ-capability limitations are discussed.

1. Types of CDMA Systems

Code-division multiple-access (CDMA) systems can be classified into major groups. The first major categorization pertains to the effect system accesses have on each other. Ideally, one desires an orthogonal relationship. That is, there is no additive interaction between system users. Such systems have been occasionally discussed in the literature from an academic point of view. Practically speaking, they require

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*References are listed at the end of the report.

excessive network-timing accuracy* to obtain the desired approximation to orthogonality. Consequently, orthogonal CDMA will receive no further consideration.

The other major CDMA category is sometimes known as quasi-orthogonal CDMA. In these systems, the effect of other accesses can be modeled by bandlimited white noise of the same bandwidth and power. The total power of the other accesses is reduced by the processing gain, but it is finite and nonzero as contrasted with the situation for orthogonal CDMA.† All practical CDMA systems fall within the category of quasi-orthogonal multiple access. We choose to call this category spread-spectrum multiple access (SSMA).

SSMA may be further broken down into several categories. One such dichotomization is based on whether the spread-spectrum modulation is coherent or noncoherent. Coherent SSMA (which might be designated SSMAC) is basically a phase-modulation approach. It is the most common form of spread spectrum. In its most common form it is sometimes known as direct-sequence spread spectrum. Phase-chipping spread spectrum

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*Typically, the timing accuracies required are far in excess of those required by a TDMA system. For a typical TDMA system a timing accuracy of 100 nanoseconds is adequate. However, for a near-orthogonal CDMA system of similar capability, timing accuracies better than 1 nanosecond would be required.

†For this result to hold, the codes must be independent of one another over the bit duration time. This is not normally a problem when pseudo-random sequences of reasonable lengths (much longer than a data bit) are employed. Much work has been performed to determine the cross-correlation between different maximal-length (M) sequences.[2] Tables have been generated listing groups of M sequences with minimal cross-correlation. Practically, these interesting theoretical results have little meaning since they refer to the cross-correlation over a complete code period. The correlation performed by the receiver is over the much shorter bit duration. Fortunately, for this situation, virtually any other M sequence can be regarded as independent. Thus, the quasi-orthogonal assumption does not require elaborate code selection.

(PCMA) is probably the best description. Noncoherent SSMA (which might be designated SSMAN) is basically a frequency-modulation approach. Frequency-hopping spread spectrum (FHMA) is probably the best and most common description of such systems. While there are significant performance differences between SSMAC and SSMAN, these differences are more related to AJ performance and initial acquisition time than to multiple-access performance. Typically, the results will be derived for the SSMAC case but will apply equally to SSMAN.

Another dichotomy (independent of the above one) is based on the envelope function associated with the spread-spectrum modulation. Normally one has a constant envelope and obtains the above-described spread-spectrum formats. However, it is possible to employ on-off amplitude modulation with the pseudo-random angle modulation. If this is done with coherent spread spectrum, the result is known as asynchronous time-gated spread spectrum. If it is done with noncoherent spread spectrum, the result is known as pulse-address multiple access (PAMA). The various random-access discrete-address (RADA) systems are examples of this format. The primary advantage of the on-off modulation is that it permits improved performance in the presence of a large dynamic range in signal powers. The disadvantage is that very poor utilization is made of the power capabilities of the satellite transponder since it is not occupied for a very significant fraction of the time.[1] In general, it has proven more efficient to accept the power-control losses than to attempt to design an inefficient system that is unaffected by power-level discrepancies. Consequently, consideration will be restricted to constant-envelope spread-spectrum multiple access.

Coherent spread spectrum with constant envelope can be further categorized. Perhaps the most common form is known as <u>direct sequence</u> SSMA and utilizes the unfiltered output of a biphase or quadriphase modulator driven by a PN sequence. The NOMAC (noise-modulation and

correlation) type of SSMA is similar except that a small section of the spectrum of a direct-sequence signal is selectively filtered to produce a random waveform (similar to the approach used in Hewlett-Packard's pseudo-random noise generator) that possesses a nonconstant random envelope as well as a random phase. It is necessary to run the direct-sequence portion of the system much faster than the resulting "spread" bandwidth. Consequently, there are some implementation problems but the resulting waveform is more noise-like than for the direct-sequence case. Equipment simplicity favors the direct-sequence approach, which is in more common use.

Figure 2 illustrates the relationship between the different types of code-division multiple access.

2. Multiple-Access Capability

The following four subsections analyze the multiple-access capability of the SSMA form of CDMA. The first three sub-sections are fundamental but approximate analyses that apply to the cases of: (1) identical accesses, (2) differences in receive G/T, and (3) differences in receive G/T and data rates. The fourth subsection is a precise analysis of the performance of a direct-sequence SSMA system. It precisely defines the bandwidth/processing gain relationship and determines the variability in the resulting performance as a function of system timing. All analyses assume an ideal, AGC type transponder.

a. Identical Accesses--RF Bandwidth and Bandwidth Utilization

The purpose of this subsection is to present fundamental limits on the performance of SSMA systems and the derivation of these limits.

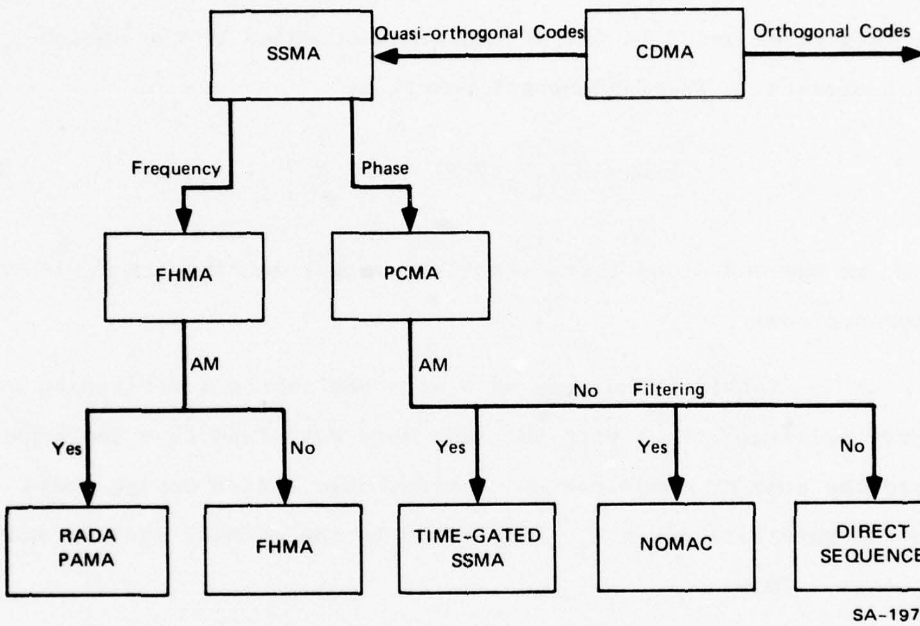

FIGURE 2 ILLUSTRATION OF RELATIONSHIPS AMONG DIFFERENT TYPES OF CODE-DIVISION MULTIPLE ACCESS

A commonly used and good measure of the error-rate capability of a digital spread-spectrum modem is the signal-to-noise power ratio (SNR) at the correlator (matched filter or bit detector)

output.* The SNR is given as the ratio of the desired signal power P_s to the sum of gaussian receiver noise plus interference effects. For the case of spread-spectrum multiple-access systems the interference is due to the other accesses that are present. The access noise at the correlator output is reduced by the processing gain of the spread-spectrum receiver. The processing gain is given approximately by b/W where b is the equivalent noise bandwidth of the matched filter (which is determined by the data rate) and W is the RF bandwidth occupied by the spread-spectrum signals.† The fundamental result is

$$SNR = P_s / \left[P_I (b/W) + b \cdot N_o \right] \tag{1}$$

where N_o is the one-sided thermal-noise spectral density and P_I is the interference power.

Consider the case of N such SSMA users (overlapping n frequency and time), each with the same data rate (and thus the same "b") and the same RF bandwidth W. A reasonable system design would divide the satellite power P_T (referenced to the ground) equally among the N users. Thus,

$$P_s = P_T/N \tag{2}$$

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*This SNR precisely specifies the error rate for the case of an additive gaussian noise channel. For additive disturbances with other statistical characterizations the SNR remains a good if not precise measure of error rate. Precise evaluation requires determination of all higher-order moments. Typically, precise and approximate evaluations are quite close except for the case of exceptionally lower error rates (e.g., 10^{-13}), which are not normally required.

†The processing gain may differ from the above value by a factor of 2, depending on the details of the matched filter and RF spectral shapes. The value of b/W is presented as a nominal value that may be used to derive the basic limitations.

and

$$P_I = \left(1 - \frac{1}{N}\right)P_T \quad . \tag{3}$$

Substitution of Eqs. (2) and (3) in Eq. (1) yields

$$SNR = \frac{P_T}{\left[N\ (1 - 1/N)P_T \cdot (b/W) + bN_o\right]} = \frac{1}{\left[N\ (1 - 1/N)b/W + bN_o/P_T\right]} \quad . \tag{4}$$

To achieve the desired system operation it is required that SNR = X where X is specified to yield the desired error rate. Typically, X = 10 is a good value to yield an error rate of 10^{-5}. If b, P_T/N_o (the down-link capacity quotient),* and X are specified, then it is possible to find W as a function of the number of users. Equation (4) has been programmed and run for several values of X, b, and P_T/N_o.† Figure 3 plots the results, which are the required RF bandwidth W as a function of the number of the system accesses N. The values have been plotted for down-link capacity quotients of 70 and 80 dB (Hz), assuming all accesses have the same data rate (100 kbps), power, and required E/N_o (10 dB). For the case of a 70-dB (Hz) capacity quotient the maximum number of users that could be supported on the basis of power arguments only is 10. Operation under this circumstance requires an infinite RF bandwidth. If the number of users is decreased to 9, then the RF bandwidth may be reduced to 80 MHz. The case of an 80-dB (Hz) capacity quotient is also plotted. Sufficient power is available to support the maximum number (24) of accesses considered. Less bandwidth is required for the higher-capacity-quotient case since the desired E/N_o can be obtained with a higher level of access noise (after processing gain).

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*The capacity quotient is a measure of a link's ability to convey information. For example, a quotient of 80 dB(Hz) permits 10Mbps to be transmitted with an-energy-per-bit-to-noise-power-spectral-density ratio of 10 dB.

†This program (SSMA) is presented in Appendix A, Figure A-1.

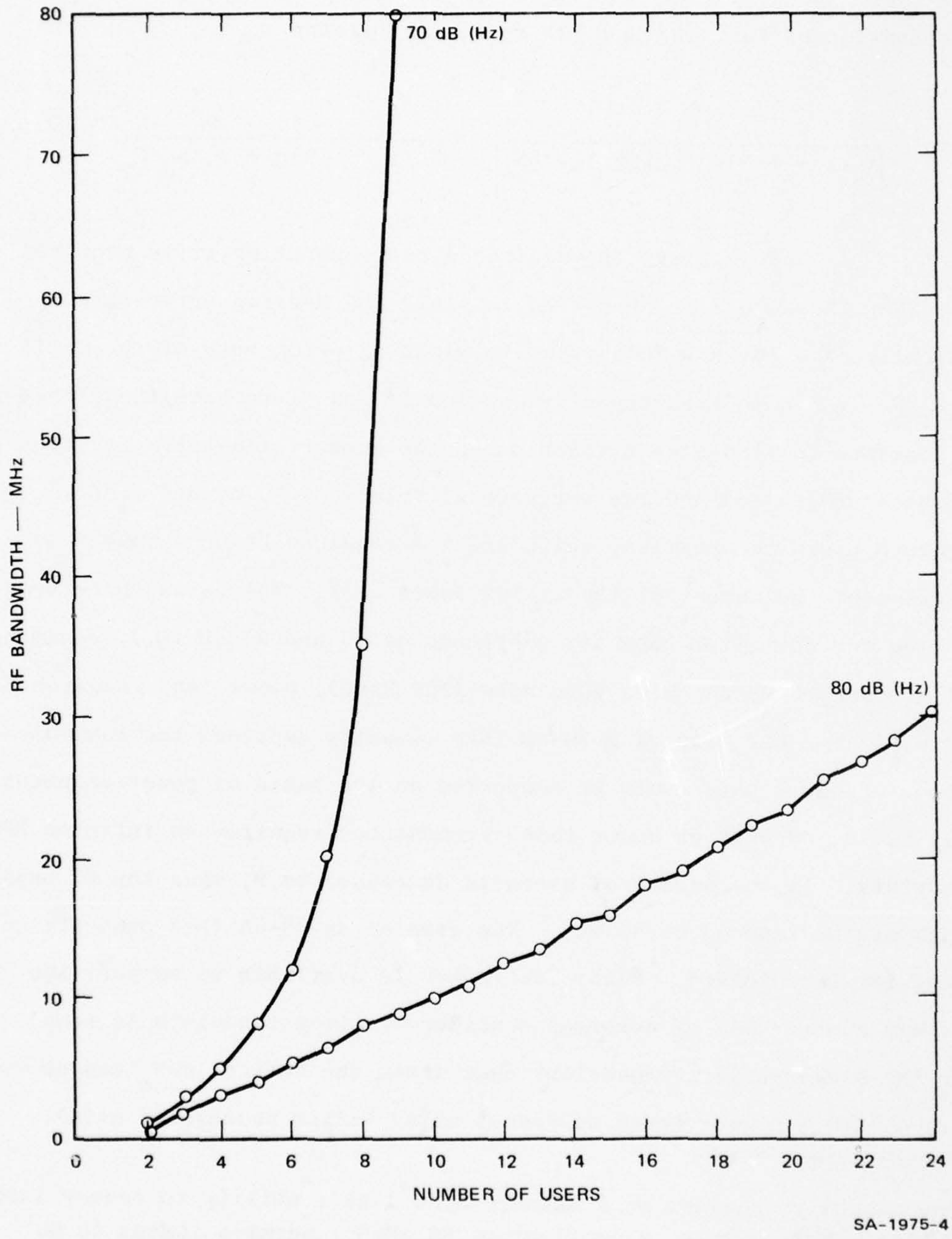

FIGURE 3 RF BANDWIDTH AS A FUNCTION OF THE NUMBER OF IDENTICAL ACCESSES

Equation (4) can be rewritten as

$$W = b X (N - 1)/[1 - NbX/Q] \tag{5}$$

where $Q = P_T/N_o$. Equation (5) states that if $Q/NbX \gg 1$, then the required RF bandwidth is a linear function of the number of accesses. However, as the number of accesses is increased, the inequality will no longer be satisfied and the denominator of Eq. (5) must be considered. It is the denominator that causes the knee in the upper curve of Figure 3. Physically, the knee results from the following situation. When the SSMA system approaches the point of being power-limited, then it is necessary to use larger and larger bandwidth to render the processed access noise smaller and smaller. In the limit, as the power limit (assuming an orthogonal access technique) is reached, an infinite RF bandwidth is required to reduce the access noise to zero.

Equation (4) can be used to determine the required RF bandwidth to approach orthogonal performance for the case of a large number of accesses. Performance within 1 dB of orthogonal capability will be achieved when the first term in the denominator is less than 0.25 of the value of the second term. This is equivalent to the requirement that $W > 4Q$. Table 1 lists the degradation as a function of W/Q and the bandwidth utilization (BU). The latter quantity is based on a desired (with ∞ bandwidth) SNR of 10.

Program SSMA also evaluates the bandwidth utilization as a function of the number of users. Inputs are the data rates, capacity quotient, and desired E/N_o. Table 2 presents results obtained for E/N_o = 10 dB, data rate = 100 kbps, and a desired number of users of 24. Note that for the case Q = 70 dB (Hz) it is possible to support only 10 users, while for the other cases it is possible to support the full 24. The general expression for bandwidth utilization is

Table 1

FINITE-BANDWIDTH DEGRADATION

W/Q	BU (percent)	Degradation (dB)
4	2.5	1.0
9	1.1	0.45
19	0.53	0.22
40	0.25	0.1

Table 2

BANDWIDTH UTILIZATION

Q [dB (Hz)]	Number of Users	Utilization (percent)
70	10	0
80	24	7.9
90	24	10.2
100	24	10.4

$$BU = (Nb/W) = Nb(Q - NbX)/[bQX(N - 1)] \quad . \tag{6}$$

For large N this becomes

$$BU \approx (Q - NbX)/QX \quad . \tag{7}$$

From Eq. (7) one can see that the bandwidth utilization approaches zero as the number of users approaches the power limitation on the number of accesses. This is the case since the required RF bandwidth approaches infinity.

b. Effect of Unequal Power and Bit Rates on Power Efficiency and Power Control

Real operating conditions for satellite communication systems involve a mix of user types. That is, in general, users will have different data rates, transmitter ERP's, and receiver G/T's. Thus, imbalances in power and spectral densities of different accesses will exist. It is important to determine the impact of these imbalances on the performance of the different multiple-access techniques so that truly meaningful results are obtained. The purpose of this subsection is to accomplish this for spread-spectrum multiple access.

There is good reason to believe that SSMA may be more vulnerable to link imbalances than FDMA or TDMA. The latter is a truly orthogonal* multiple-access technique, while FDMA is approximately orthogonal--the difference being due to the presence of intermodulation cross-products and the power-sharing and suppression effects. However, SSMA is a quasi-orthogonal multiple-access technique. That is, the presence of the other accesses appears as an additive "noise" power that is in proportion to their power and reduced by the (finite) processing gain. Thus, in contrast to TDMA, the presence of the other access and their power levels has a direct effect on the correlator output SNR. Consequently, SSMA will be degraded more (than TDMA) by the high-spectral-density signals necessary to overcome the low G/T of the intended receiving terminals. In this note we analyze this effect to determine its significance.

Computer programs were developed to assess the performance of SSMA under the above circumstances.† The programs were developed for

*By orthogonal we mean that the accessing signals are independent of one another. That is, the presence and character of one access has no effect on the error rate of the other accesses.

†The programs are given in Figures A-2, A-3, and A-4.

the special case of two accesses. Each access occupied the full RF bandwidth but the spectral densities differed due to a substantial power difference.

Program SSPC assumes that there are two accesses at the same data rate and that the E/N_o required at each is X and Y, respectively. The thermal noise present at each is P1 and P2, respectively. The processing gain is given by 1/K. The first portion of the program evaluates the required power (at the ground) in each signal A1 and A2, respectively, when the access noise is negligible--e.g., in the case of infinite RF bandwidth. The values A1 and A2 are considered to be an output under the case of orthogonal power control. The second portion of the program evaluates the required powers C1 and C2 when the access noise due to finite RF bandwidth is properly included in the analysis. These results are obtained under what is considered to be quasi-orthogonal power control. Greater power is required in this case to combat the access noise. The multiple-access loss is measured by calculating the ratio (C1 + C2)/(A1 + A2).

Program SSPC1 assumes that there are two accesses at the same data rate, that the E/N_o required at each is X (with infinite processing gain), and that processing gain is given by 1/K. The thermal-noise power present at each is P1 and P2, respectively. The first portion of the program evaluates the required power (at the ground) in each signal A1 and A2, respectively, when the access noise is negligible. The second portion of the program assumes that the total power remains constant at A1 + A2 but is redistributed as B1 and B2 (where B1 + B2 = A1 + A2), reflecting the impact of access noise. Since the total power remains fixed, the ratio of energy per bit to noise-power spectral density (E/N_o) at each receiver (assumed to be the same) must degrade. The resulting E/N_o is printed and the difference with respect to X is printed as the access loss.

Program SSPC2 generalizes program SSPC to include the case of different bit rates for each of the two accesses. Since bit rates are specified, noise-power spectral densities rather than noise powers are specified for each access.

In summary, Programs SSPC and SSPC2 evaluate the increase in satellite power required to maintain the same E/N_o (at both terminals) with spread-spectrum multiple access that is possible with an orthogonal multiple-access technique. This performance loss is evaluated under the assumption of optimized power control for both circumstances. Program SSPC1 assumes that the satellite power remains fixed and that the E/N_o (the same for both terminals) is degraded by the access noise. The program evaluates the E/N_o and the loss in it with respect to that obtainable with an orthogonal multiple access. These values are determined under the assumption of optimized power control.

SSPC was run for several cases where the desired E/N_o for both accesses was 10. Table 3 presents the additional satellite power required as a function of the processing gain and difference in the G/T between the two terminals.

Table 3

ACCESS LOSS AS A FUNCTION OF PROCESSING GAIN AND G/T RANGE

Processing Gain (dB)	G/T Difference (dB)	Access Loss (dB)
40	27	0.004
30	23	0.043
25	23	0.132

The largest value of access loss of 0.132 dB is sufficiently small that link unbalances do not seem to be very wasteful of satellite power even with processing gains as low as 25 dB.

In an effort to find a greater access loss the following problem was postulated. Both receivers require a 10-dB E/N_o and have an RF bandwidth of 10 MHz. User 2 has a G/T 25 dB lower than User 1. However, the data rate to User 2 is 20 dB greater than to User 1. Thus, since both signals occupy the same RF bandwidth the spectral density of User 2 is 45 dB greater than for User 1. In this case, with the optimized power control (determined by Program SSPC2) the increase in satellite power is 0.46 dB. Thus, in this severe case the loss begins to approach a noticeable value.

SSPC1 has been run for several conditions to determine the decrease in E/N_o caused by access noise as a function of link imbalances. Table 4 presents this decrease as a function of processing gain and G/T difference.

Table 4

E/N_o LOSS AS A FUNCTION OF PROCESSING GAIN AND G/T RANGE

Processing Gain (dB)	G/T Difference (dB)	E/N_o Loss (dB)
20	27	0.41
20	23	0.41
25	23	0.13
30	23	0.043
40	23	0.0043

From Table 4 we see that the E/N_o loss depends solely on the processing gain. For processing gains greater than 20 dB the loss is less than 0.5 dB. Thus, the effect is not very significant.

One might conclude that neither the requirement for increased satellite power nor the E/N_o loss is a significant effect.* However, this is a degradation that exists with SSMA and not with TDMA or FDMA, in principle.† Consequently, it should be recorded since there may well be other small losses which, if totaled, yield a significant difference among multiple-access techniques.

It should be pointed out that the above results are obtained assuming optimized power control. This power control reflects the effect of access noise very significantly. For example, consider the following case run with Program SSPCI: Difference in G/T = 23 dB and processing gain = 25 dB. With an orthogonal multiple-access technique, Users 1 and 2 require 10 and 2000 units of power, respectively. With spread-spectrum multiple access the access noise causes the power to be redistributed with 66.3 and 1944 units of power to Users 1 and 2, respectively. Thus, to combat the access noise of the stronger user the power distributed to the weaker user is increased by 8.2 dB. If this were not done, the smaller user would experience an unacceptably low E/N_o (approximately 2 dB). Thus, one can see that disparities in receiver G/T's can have significant impact on the power control for SSMA techniques.

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*Note that only one loss exists. The two losses arise from different views of the same basic phenomenon. The first loss arises when one considers that one has the option of adjusting the satellite power to provide the desired service. This is the viewpoint of a satellite system designer. The second loss is seen from the viewpoint of a system user. That is, the satellite power is fixed and the E/N_o is degraded owing to the presence of the access noise.

†Adjacent channel interference effects play a similar role in FDMA. Evaluation of the significance of these effects is covered in Section III-C-2-a.

c. SSMA Bandwidth Utilization as a Function of Differences in Data Rates and Receiver Sensitivities (G/T)

Another problem of interest is the determination of the required RF bandwidth for each SSMA access when each access has a different data rate and receiver sensitivity (G/T). In this case the different RF bandwidth W_i will be required to maintain the same signal-to-noise ratio (SNR_i) at each correlator output.

For the case of an "ideal," AGC transponder,

$$SNR_i = \left(P_{si} \Big/ \sum_{j=1}^{M} P_{sj}\right) \cdot P_T \Bigg/ \left[\left(1 - P_{si} \Big/ \sum_{j=1}^{M} P_{sj}\right) P_T \left(b_i / W_i\right) + P_{ni}\right] \tag{8}$$

where M is the number of accesses, P_{si} is the power in the j^{th} access at the satellite input, P_T is the total transponder power referenced to the ground, P_{ni} is the noise power at the i^{th} correlator output, and b_i and W_i are the data rate and RF bandwidths of the i^{th} access, respectively. If the power control is effected properly,

$$P_{sj} \sim b_j / Q_j \qquad \text{for} \qquad j = 1 \text{ to } M \tag{9}$$

where $Q_j = P_T / P_{nj}$. Using Eq. (9) in Eq. (8) one has

$$SNR_i = \frac{\left(b_i / Q_i\right) \Bigg/ \left(\sum_{j=1}^{M} b_i / Q_j\right)}{\left[1 - \left(b_i / Q_i\right) \left(\sum_{i=1}^{M} b_i / Q_j\right)\right] \left(b_i / W_i\right) + \left(b_i / Q_i\right)} \ . \tag{10}$$

Solving for W_i one finds

$$W_i = \frac{Q_i\left[\sum_{j=1}^{M}\left(b_j/Q_j\right)\right] \cdot SNR_i\left[1 - \left(b_i/Q_i\right)\Big/\sum_{j=1}^{M}\left(b_j/Q_j\right)\right]}{1 - SNR_i\left[\sum_{j=1}^{M}\left(b_i/Q_j\right)\right]} \qquad i = 1 \text{ to } M \quad . \tag{11}$$

The bandwidth utilization is given by

$$BU = \left(\sum_{i=1}^{M} b_i \Big/ w_{max}\right) \tag{12}$$

where W_{max} is the maximum values of W_i.

Equations (11) and (12) have been programmed for the special case that all accesses desire the same correlator output SNR--i.e., $SNR_i = X$. Figure A-5 is a listing of Program SSW1. Program SSW1 was run for several test cases to demonstrate the range of values that might result:

- Test Case 1

  Number of accesses = 4
  Data rates: 3 at 10 kbps
  Capacity quotients: All at 60 dB(Hz)
  Signal-to-noise ratio: 10 dB

  Results

  Bandwidth utilization: 6.25%
  Maximum bandwidth: 0.8 MHz
  Three low-data-rate users at 0.8 MHz and the high data rate user at 0.6 MHz

  This result may seem somewhat surprising, but the high-data-rate user has more power and sees less interference power. Thus it requires a lower processing gain, resulting in less RF bandwidth.

- Test Case 2

  Number of accesses: 4
  Data rates: 3 at 10 kbps and 1 at 20 kbps
  Capacity quotients: 10 kbps accesses have 60 dB(Hz), 20 kbps access has 67 dB(Hz)
  Signal-to-noise ratio: 10

  Results

  Bandwidth utilization: 2.2%
  Maximum bandwidth: 2.27 MHz
  Three low-data-rate users require only 0.364 MHz and the high-data-rate user needs 2.2 MHz.

  Only a small fraction of the transponder power goes to the high-rate access, so it needs largest processing gain and bandwidth.

- Test Case 3

  Number of accesses: 4
  Data rates: 3 at 10 kbps and 1 at 20 kbps
  Capacity quotients: All accesses 60 dB(Hz)
  Signal-to-noise ratio: 19 dB

  Results

  Bandwidth utilization: 0.69%
  Maximum bandwidth: 7.2 MHz
  High-data-rate user requires only 5.4 MHz

- Test Case 4

  Number of accesses: 4
  Data rates: 3 at 10 kbps and 1 at 100 kbps
  Capacity quotients: 10 kbps accesses at 60 dB(Hz) and 100 kbps at 67 dB(Hz)
  Signal-to-noise ratio: 10 dB

  Results

  Bandwidth utilization: 4.3%
  Maximum bandwidth: 3 MHz
  Low-rate accesses require only 0.8 MHz

One concludes that the high-data-rate users do not always require the largest bandwidth. The analysis must include the receiver sensitivity (G/T) factor. Program SSW1 is a very useful tool for

determining the required RF bandwidth* and bandwidth utilization. Note that for the selected examples the bandwidth utilization is typically well below 10 percent. Requirements for a high correlator output SNR are particularly harmful to utilization. If the high-rate access also has a very-high-capacity quotient with respect to the other users, then this situation tends to lead a poor bandwidth utilization since the high-rate user requires a high processing gain.

d. A Detailed Analysis of the Multiple-Access Capability of Binary Direct-Sequence Phase-Chipping Spread-Spectrum Systems Employing Phase-Reversal Keying

Previous subsections analyzed the performance of SSMA based on the assumption that the other accesses could be treated as additive bank-limited white gaussian noise. It has been established that the results obtained with this simplified model yield the correct relative characteristics--e.g., power and bandwidth limitations. However, the absolute values need to be confirmed by an analysis employing more precise models.

The purpose of this subsection is to provide a more exact analysis of the performance of a binary, direct-sequence, phase-chipping, spread-spectrum system employing phase-reversal keying. The results are expected to fall within ±3 dB of the result presented in earlier work.[1] While any difference of this magnitude may not seem significant, an accurate analysis is of interest for at least two reasons. First, quantitatively different theoretical results will be obtained, depending on

---

*In practice, all accesses would normally be operated at the maximum bandwidth. This means that some of the users' SNR's will be better than the required value. If all accesses have the same RF bandwidth, it may be necessary to reflect access noise in the power-control algorithm. By contrast the variable RF bandwidth approach controls power solely on the basis of data rates and receiver G/T's.

assumptions. Second, in practice the performance of a phase-chipping spread-spectrum system will be degraded by a relatively large number of sources each producing a small degradation. Since a large number of "insignificant" degradations are not insignificant, it is desirable to make an accurate accounting of the degradation budget. For example, in comparing the multiple-access capability of spread-spectrum with respect to the ideal, orthogonal system, it is common to claim that for sufficiently large RF bandwidth, spread-spectrum has ideal performance. This, in fact, is not the case, as the following sample degradation budget shows.

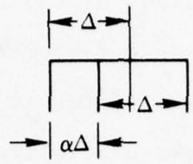
1 dB	Suppression loss
1 dB	Unbalanced power-control loss
0.5 dB	Phase-code tracking loss*
1 dB	Finite-bandwidth loss†
3.5 dB	Total *(estimated)*

Thus, for this estimated degradation budget, spread spectrum will achieve only 45 percent of the data capability of an ideal system. Clearly it is important to keep an accurate budget of "insignificant" degradations.

Video Analysis. As a preliminary step it is desirable to determine the performance on a video basis when two quasi-orthogonal codes are present. It is assumed that the two codes, $a_1(t)$ and $a_2(t)$, are clocked at the same rate, but the transitions may shift in time with respect to each other. The shift is represented by the factor α shown in Figure 4 where Δ represents the chip duration. Note that for $\alpha = 0$ and 1

*This loss assumes that a single correlator (time-shared between early and late channels) delay-lock loop is used for code tracking.

†This loss includes degradation due to wasted power (outside the passband), signal distortion due to amplitude and phase effects, and access noise from other system users.



SA-1975-5

FIGURE 4 DIAGRAM SHOWING CHIP OVERLAP

the chips of the two codes have transitions at the *same* instances. The mean correlator output produced by another quasi-orthogonal code is given by

$$\bar{z} = E\left\{\int_0^{T=N\Delta} s_1(t)\, s_2(t)\, dt\right\} \tag{13}$$

$$= E\left\{\sum_{i=1}^{N} \int_0^{\Delta} a_i s_{2i}(t)\, dt\right\} \tag{14}$$

$$= E\left\{\sum_{i=1}^{N} \left[\int_0^{\alpha\Delta} a_i b_i\, dt + \int_{\alpha\Delta}^{\Delta} a_i c_i\, dt\right]\right\} \tag{15}$$

$$= E\left\{\sum_{i=1}^{N} \left[\alpha\Delta\; a_i b_i + (1-\alpha)\Delta a_i c_i\right]\right\} \tag{16}$$

$$= \sum_{i=1}^{N} \left[\alpha\Delta\; \bar{a}_i \cdot \bar{b}_i + (1-\alpha)\Delta\; \bar{a}_i \cdot \bar{c}_i\right] = 0 \tag{17}$$

where $a_i = \{+1, -1\}$ represents the values of the N phase chips of the desired phase code. The parameters b_i and c_i represent the leading and

lagging overlapping phase-chip values of the second code. These chips also have range $\{+1, -1\}$ or any scale factor times this range. For the purposes of this analysis the scale factor is assumed unity. The binary random variables a_i, b_i, and c_i are statistically independent.

Consider now the variance of the correlator output

$$\mathrm{Var}\ \{z\} = E\left\{\sum_{i=1}^{N}\left[\alpha\Delta a_i b_i + (1-\alpha)\Delta a_i c_i\right]\right\}^2 \quad . \tag{18}$$

Let $\gamma_i = \alpha\Delta a_i b_i + (1-\alpha)\Delta a_i c_i$. Then Eq. (18) may be rewritten as

$$\mathrm{Var}\ \{z\} = E\left\{\sum_{i,j=1}^{N}\gamma_i\gamma_j\right\}$$

$$= E\left\{\sum_{i,j=1}^{N}\left[\alpha^2\Delta^2\ a_i a_j b_i b_j + (1-\alpha)^2\ a_i a_j c_i c_j\right.\right.$$

$$\left.\left. + \alpha\Delta(1-\alpha)\Delta\ a_i b_i a_j c_j + (1-\alpha)\alpha\Delta\ a_i c_i a_j b_j\right]\right\} \quad . \tag{19}$$

Interchanging the order of the expectation operator and the summation, and taking advantage of symmetry, one finds that

$$\mathrm{Var}\ \{z\} = \sum_{i,j=1}^{N}\left[\alpha^2\Delta^2\ \overline{a_i a_j} \cdot \overline{b_i b_j} + (1-\alpha)^2\Delta^2\ \overline{a_i a_j} \cdot \overline{c_i c_j}\right.$$

$$\left. + 2\alpha\Delta(1-\alpha)\Delta\ \overline{a_i a_j} \cdot \overline{b_i c_j}\ \right] \tag{20}$$

where the independence of a_i, b_i, and c_i has been used.

The phase codes may be expected to have desirable autocorrelation functions. Consequently, Eq. (20) may be simplified to

$$\mathrm{Var}\,\{z\} = \sum_{i=1}^{N} \left[\alpha^2 \Delta^2 \, \overline{a_i^2} \cdot \overline{b_i^2} + (1 - \alpha)^2 \Delta^2 \, \overline{a_i^2} \cdot \overline{c_i^2} \right.$$

$$\left. + \, 2\alpha(1 - \alpha)\Delta^2 \sum_{i,j=1}^{N} \left[\overline{a_i a_j} \cdot \overline{b_i c_j} \right] \right] \quad . \tag{21}$$

Using the facts that

$$\overline{a_i^2} = \overline{b_i^2} = \overline{c_i^2} = 1 \qquad \text{and} \qquad c_j = b_{j+1} \quad ,$$

one notes that $\overline{a_i a_j} = 0$ unless $i = j$; however, $\overline{b_i c_j} = 0$ for $i = j$. Similarly, $\overline{b_i c_j} = 0$ unless $j = i + 1$; however, then $a_i a_j = 0$. Consequently, Eq. (21) reduces to

$$\mathrm{Var}\,\{z\} = N\left[\alpha^2 + (1 - \alpha)^2\right]\Delta^2 \quad . \tag{22}$$

Thus, the cross-channel interference produced by another user is a function $g(\alpha)$ of the overlap α:

$$g(\alpha) = \alpha^2 + (1 - \alpha)^2 \tag{23}$$

Figure 5 illustrates this symmetrical function. Note that the minimum cross-channel interference occurs for a half-chip overlap, and this interference level is 3 dB better than the worst case of simultaneous chip transitions.

The dependence of cross-channel interference on chip overlap interval may seem surprising to those familiar with the usual

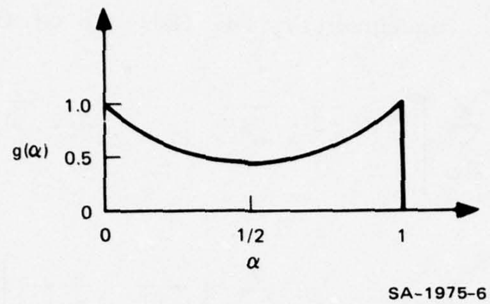


FIGURE 5 EFFECT OF OVERLAP FACTOR α ON CROSS-CHANNEL-INTERFERENCE POWER

processing-gain approach to the analysis of spread-spectrum systems. The processing-gain approach utilizes a convolution theorem that strictly applies only for statistically independent processes applied to the correlation multiplier. However, two quasi-orthogonal phase codes clocked at the same rate are not truly statistically independent. Consequently, the processing-gain approach may be expected to be in error and insensitive to the degree of chip overlap.

Figure 5 illustrates the best and worst cases. The average value is given by

$$E\left\{g(\alpha)\right\} = \int_0^1 p(\alpha)g(\alpha)\,d\alpha = \int_0^1 \left\{\alpha^2 + (1-\alpha)^2\right\} d\alpha = 2/3 \quad . \tag{24}$$

Thus, on the average, performance will be 1.75 dB better than the lower bound based on synchronous transition. However, in practice when only two accesses are present there will be relatively long periods of time when the lower bound is achieved. This periodic fluctuation occurs due to slight differences in the clocking frequencies of the two sequences.

Bandpass Analysis. The above result was derived on a video basis. It is natural to wonder if the same sensitivity to chip overlap occurs on a bandpass basis. The subsequent analysis shows that indeed the same result does hold.

Consider the variance of the correlator output produced by cross-channel interference

$$\mathrm{Var}\,\{z\} = E\left\{\int_0^{N\Delta} s_1(t)\, s_2(t)\, dt\right\}^2$$

$$= E\left\{\sum_{i=1}^{N}\left[\int_0^{\alpha\Delta} a_i b_i \cos\,\theta_1 \cos\,\theta_2 dt + \int_{\alpha\Delta}^{\Delta} a_i c_i \cos\,\theta_1 \cos\,\theta_2 dt\right]\right\} \qquad (25)$$

where $\theta_1 = \omega_1 t + \varphi_1$ and $\theta_2 = \omega_2 t + \varphi_2$. Equation (25) reduces to

$$\mathrm{Var}\,\{z\} = E\left\{\sum_{i=1}^{N}\left[a_i b_i\, \eta + a_i c_i\, \xi\right]\right\}^2 \qquad (26)$$

where

$$\eta \triangleq \int_0^{\alpha\Delta} \psi(t) dt$$

$$\xi \triangleq \int_{\alpha\Delta}^{\Delta} \psi(t) dt$$

where

$$\psi(t) \triangleq \cos\,\theta_1\, \cos\,\theta_2.$$

Thus, again,

$$\mathrm{Var}\,\{z\} = E\left\{\sum_{i,j=1}^{N} \gamma_i \gamma_j\right\} \tag{27}$$

where $\gamma_i \stackrel{\Delta}{=} a_i b_i \eta + a_i c_i \xi$ is now a more complicated form. Rewriting Eq. (27) in terms of this identity, one has

$$\mathrm{Var}\,\{z\} = E\left\{\sum_{i,j}^{N} \left[a_i a_j b_i b_j \eta^2 + a_i a_j c_i c_j \xi^2 + 2a_i b_i a_j c_j \eta\xi\right]\right\}$$

$$= \sum_{i,j}^{N} \left[\overline{a_i a_j}\;\overline{b_i b_j}\;\overline{\eta^2} + \overline{a_i a_j} \cdot \overline{c_i c_j}\;\overline{\xi^2} + 2\overline{a_i a_j} \cdot \overline{c_i b_j}\;\overline{\xi\eta}\right] \quad . \tag{28}$$

Utilizing the same arguments employed on the video basis, one finds that

$$\mathrm{Var}\,\{z\} = \sum_{i=1}^{N} \left[\overline{a_i^2}\;\overline{b_i^2}\;\overline{\eta^2} + \overline{a_i^2}\;\overline{c_i^2}\;\overline{\xi^2}\right] + 2 \sum_{i,j=1}^{N} \left[\overline{a_i a_j} \cdot \overline{b_i c_j} \cdot \overline{\xi\eta}\right]$$

$$= N\left[\overline{\eta^2} + \overline{\xi^2}\right] \quad . \tag{29}$$

The next step is to evaluate the two terms within the bracket. Consider the first term:

$$\overline{\eta^2} = \overline{\int_0^{\alpha\Delta} \cos(\omega_1 t + \varphi_1)\cos(\omega_2 t + \varphi_2)dt \cdot \int_0^{\alpha\Delta} \cos(\omega_1 \nu + \varphi_1)\cos(\omega_2 \nu + \varphi_2)d\nu} \tag{30}$$

$$= \int^{\alpha}\!\!\int_0^{\Delta} \overline{\cos(\omega_1 t + \varphi_1)\cos(\omega_1 \nu + \varphi_1)\cos(\omega_2 t + \varphi_2)\cos(\omega_2 \nu + \varphi_2)}\,dt\,d\nu \tag{31}$$

$$\eta^2 = 1/2 \int^{\alpha}\!\!\int_0^{\Delta} \left[\cos \omega_1(t - \nu) + \cos\overline{\{\omega_1(t + \nu) + 2\varphi_1\}}\right] \times \left[\cos \omega_2(t - \nu)\right.$$

$$\left. + \cos\overline{\{\omega_2(t + \nu) + 2\varphi_2\}}\right] dt\, d\nu \tag{32}$$

$$= 1/2 \int^{\alpha}\!\!\int_0^{\Delta} \cos \omega_1 s \cdot \cos \omega_2 s \, d\nu\, dt \tag{33}$$

where $s \overset{\Delta}{=} t - \nu$. Evaluating the integral with respect to s and t yields

$$\overline{\eta^2} = 1/2\ (-1) \int_0^{\alpha\Delta}\!\!\int_t^{t-\alpha\Delta} \left[\cos \omega_1 s \times \cos \omega_2 s\right] ds\, dt \tag{34}$$

$$= -\ 1/2 \int_0^{\alpha\Delta} dt \cdot 1/2 \int_t^{t-\alpha\Delta} \left[\cos\ (\omega_1 - \omega_2)s + \cos(\omega_1 + \omega_2)s\right] ds \tag{35}$$

$$= -\ 1/4 \int_0^{\alpha\Delta} dt \left\{ \frac{1}{(\omega_1 - \omega_2)} \sin(\omega_1 - \omega_2)s \,\Bigg|_t^{t-\alpha\Delta}\right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)} \sin(\omega_1 + \omega_2)s \,\Bigg|_t^{t-\alpha\Delta} \right\} \tag{36}$$

$$= -\ 1/4 \int_0^{\alpha\Delta} \left\{ \frac{1}{(\omega_1 - \omega_2)} \left[\sin(\omega_1 - \omega_2)(t - \alpha\Delta) - \sin(\omega_1 - \omega_2)t\right]\right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)} \left[\sin(\omega_1 + \omega_2)(t - \alpha\Delta) - \sin(\omega_1 + \omega_2)t\right] \right\} \tag{37}$$

and

$$\overline{\eta^2} = -1/4\left\{\frac{(-1)}{(\omega_1 - \omega_2)}\left[\cos(\omega_1 - \omega_2)(t - \alpha\Delta)\Big|_0^{\alpha\Delta} - \cos(\omega_1 - \omega_2)t\Big|_0^{\alpha\Delta}\right]\right.$$

$$\left. + \frac{(-1)}{(\omega_1 + \omega_2)^2}\left[\cos(\omega_1 + \omega_2)(t - \alpha\Delta)\Big|_0^{\alpha\Delta} - \cos(\omega_1 + \omega_2)t\Big|_0^{\alpha\Delta}\right]\right\} \tag{38}$$

$$= -1/4\left\{\frac{(-1)}{(\omega_1 - \omega_2)^2}\left[\left[1 - \cos(\omega_1 - \omega_2)(+\alpha\Delta)\right] - \left[\cos(\omega_1 - \omega_2)\alpha\Delta - 1\right]\right]\right.$$

$$\left. + \frac{(-1)}{(\omega_1 + \omega_2)^2}\left[\left[1 - \cos(\omega_1 + \omega_2)(+\alpha\Delta)\right] - \left[\cos(\omega_1 + \omega_2)\alpha\Delta - 1\right]\right]\right\} \tag{39}$$

$$\overline{\eta^2} = \frac{-1}{4}\left\{\frac{(-1)}{(\omega_1 - \omega_2)^2}\cdot 2\left[1 - \cos(\omega_1 - \omega_2)\alpha\Delta\right]\right.$$

$$\left. + \frac{(-1)}{(\omega_1 + \omega_2)^2}\cdot 2\left[1 - \cos(\omega_1 + \omega_2)\alpha\Delta\right]\right\} \tag{40}$$

$$\overline{\eta^2} = \frac{1}{2}\left\{\frac{1}{(\omega_1 - \omega_2)^2}\left[1 - \cos(\omega_1 - \omega_2)\alpha\Delta\right]\right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)^2}\left[1 - \cos(\omega_1 + \omega_2)\alpha\Delta\right]\right\} \quad . \tag{41}$$

Similarly, we evaluate the second term:

$$\overline{\xi^2} = \frac{1}{2}\int_{\alpha\Delta}^{\Delta}\int_{t-\alpha\Delta}^{t-\Delta}[\cos\omega_1 s \cdot \cos\omega_2 s]\, ds dt\ (-1) \tag{42}$$

$$\overline{\xi^2} = \frac{(-1)}{2} \int_{\alpha\Delta}^{\Delta} dt \, \frac{1}{2} \int_{t-\alpha\Delta}^{t-\Delta} \left[\cos(\omega_1 - \omega_2)s + \cos(\omega_1 + \omega_2)s\right] ds \tag{43}$$

$$= -\frac{1}{4} \int_{\alpha\Delta}^{\Delta} dt \left\{ \frac{1}{(\omega_1 - \omega_2)} \sin(\omega_1 - \omega_2)s \Bigg|_{t-\alpha\Delta}^{t-\Delta} \right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)} \sin(\omega_1 + \omega_2)s \Bigg|_{t-\alpha\Delta}^{t-\Delta} \right\} \tag{44}$$

$$= \left(-\frac{1}{4}\right) \int_{\alpha\Delta}^{\Delta} dt \left\{ \frac{1}{(\omega_1 - \omega_2)} \left[\sin(\omega_1 - \omega_2)(t - \Delta) - \sin(\omega_1 - \omega_2)(t - \alpha\Delta)\right] \right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)} \left[\sin(\omega_1 + \omega_2)(t - \Delta) - \sin(\omega_1 + \omega_2)(t - \alpha\Delta)\right] \right\}$$

$$= \left(\frac{-1}{4}\right) \left\{ \frac{(-1)}{(\omega_1 - \omega_2)^2} \left[\cos(\omega_1 - \omega_2)(t - \Delta) \Bigg|_{\alpha\Delta}^{\Delta} - \cos(\omega_1 - \omega_2)(t - \alpha\Delta) \Bigg|_{\alpha\Delta}^{\Delta}\right] \right.$$

$$\left. + \frac{(-1)}{(\omega_1 + \omega_2)^2} \left[\cos(\omega_1 + \omega_2)(t - \Delta) \Bigg|_{\alpha\Delta}^{\Delta} - \cos(\omega_1 + \omega_2)(t - \alpha\Delta) \Bigg|_{\alpha\Delta}^{\Delta}\right] \right\}$$

$$\overline{\xi^2} = \left(\frac{1}{4}\right)\left\{\frac{1}{(\omega_1 - \omega_2)^2}\left[\left[1 - \cos(\omega_1 - \omega_2)(1 - \alpha)\Delta\right]\right.\right.$$

$$\left.\left. - \left[\cos\{(\omega_1 - \omega_2)(1 - \alpha)\Delta\} - 1\right]\right]\right. \tag{45}$$

$$+ \frac{1}{(\omega_1 + \omega_2)^2}\left[\left[1 - \cos(\omega_1 + \omega_2)(1 - \alpha)\Delta\right]\right.$$

$$\left.\left. - \left[\cos\{(\omega_1 + \omega_2)(1 - \alpha)\Delta\} - 1\right]\right]\right\} \tag{46}$$

$$= \frac{1}{4}\left\{\frac{2}{(\omega_1 - \omega_2)^2}\left[1 - \cos(\omega_1 - \omega_2)(1 - \alpha)\Delta\right]\right.$$

$$\left. + \frac{2}{(\omega_1 + \omega_2)^2}\left[1 - \cos(\omega_1 + \omega_2)(1 - \alpha)\Delta\right]\right\} \tag{47}$$

$$\overline{\xi^2} = \frac{1}{2}\left\{\frac{1}{(\omega_1 - \omega_2)^2}\left[1 - \cos(\omega_1 - \omega_2)(1 - \alpha)\Delta\right]\right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)^2}\left[1 - \cos(\omega_1 + \omega_2)(1 - \alpha)\Delta\right]\right\} \quad . \tag{48}$$

Now combine the first and second terms by using Eqs. (41) and (48):

$$\overline{\eta^2} + \overline{\xi^2} = \frac{1}{2}\left[\frac{1}{(\omega_1 - \omega_2)^2}\left[2 - \cos\underbrace{(\omega_1 - \omega_2)}_{r}\alpha\Delta - \cos\underbrace{(\omega_1 - \omega_2)}_{r}(1 - \alpha)\Delta\right]\right.$$

$$\left. + \frac{1}{(\omega_1 + \omega_2)^2}\left[2 - \cos(\omega_1 + \omega_2)\alpha\Delta - \cos(\omega_1 + \omega_2)(1 - \alpha)\Delta\right]\right] \tag{49}$$

$$\overline{\eta^2} + \overline{\xi^2} = \frac{1}{2}\left\{\frac{1}{(\omega_1 - \omega_2)^2}\left[2 - 2\cos\left\{\frac{(\omega_1 - \omega_2)\Delta}{2}\right\}\cos\left\{(\omega_1 - \omega_2)\alpha\Delta\right.\right.\right.$$

$$\left.\left. - \frac{(\omega_1 - \omega_2)\Delta}{2}\right\}\right]$$

$$+ \frac{1}{(\omega_1 + \omega_2)^2}\left[2 - 2\cos\left\{\frac{(\omega_1 + \omega_2)\Delta}{2}\right\}\cos\left\{(\omega_1 + \omega_2)\alpha\Delta\right.\right.$$

$$\left.\left.\left. - \frac{(\omega_1 + \omega_2)\Delta}{2}\right\}\right]\right\} \qquad (50)$$

$$= \frac{1}{(\omega_1 - \omega_2)^2}\left[1 - \cos\frac{(\omega_1 - \omega_2)\Delta}{2}\cos\left\{(\omega_1 - \omega_2)\alpha\Delta - \frac{(\omega_1 - \omega_2)\Delta}{2}\right\}\right]$$

$$+ \frac{1}{(\omega_1 + \omega_2)^2}\left[1 - \cos\frac{(\omega_1 + \omega_2)\Delta}{2}\cos\left\{(\omega_1 + \omega_2)\alpha\Delta\right.\right.$$

$$\left.\left. - \frac{(\omega_1 + \omega_2)\Delta}{2}\right\}\right] \quad . \qquad (51)$$

In theory and in practice the last term can be neglected for any reasonable purposes. Thus,

$$\overline{\eta^2} + \overline{\xi^2} \approx \frac{1}{(\omega_1 - \omega_2)^2}\left[1 - \cos\frac{(\omega_1 - \omega_2)\Delta}{2}\cos\left\{(\omega_1 - \omega_2)\alpha\Delta\right.\right.$$

$$\left.\left. - \frac{(\omega_1 - \omega_2)\Delta}{2}\right\}\right] \quad . \qquad (52)$$

Consider the effect of α on Eq. (52). Typically, $(\omega_1 - \omega_2)/2\pi \ll 1/\Delta$--i.e., vertical stacking is employed. Consequently, $(\omega_1 - \omega_2)\Delta/2 \ll \pi/2$, which implies that $\cos\{(\omega_1 - \omega_2)\Delta/2\} > 0$. Thus, if one desires to minimize $\overline{\eta^2} + \overline{\xi^2}$, one must maximize

$$\cos\left\{(\omega_1 - \omega_2)\alpha\Delta - \frac{(\omega_1 - \omega_2)\Delta}{2}\right\} \quad .$$

Obviously the latter occurs for $\alpha = 1/2$. Thus, as for the video case, a minimum of cross-channel interference occurs for a half-chip overlap.

Again for $\alpha = 1$ or 0 a maximum occurs. The ratio between the minimum and maximum cross-channel interference is given by

$$\frac{\mathrm{Var}\{z\}_{min}}{\mathrm{Var}\{z\}_{max}} = \left\{1 - \cos\left[\frac{\omega_1 - \omega_2\ \Delta}{2}\right]\right\} \Big/ \left\{1 - \cos^2\left[\frac{(\omega_1 - \omega_2)\Delta}{2}\right]\right\}$$

$$\approx \left\{\frac{(\omega_1 - \omega_2)^2\Delta^2}{8}\right\} \Big/ \left\{\frac{2(\omega_1 - \omega_2)^2\Delta^2}{8}\right\} \approx \frac{1}{2} \quad . \tag{53}$$

Thus, a 3-dB improvement also occurs in the bandpass analysis with a half-chip overlap.

Let us now determine the performance of the spread-spectrum system. An important first step is to rewrite Eq. (52) under the condition that $(\omega_1 - \omega_2)/2\pi \ll 1/\Delta$. After numerous steps it may be shown that

$$\overline{\xi^2} + \overline{\eta^2} \approx \frac{\Delta^2}{4}\left[(1 - \alpha)^2 + \alpha^2\right] \quad . \tag{54}$$

Consider now the performance measure d^2 specified by

$$d^2 = \bar{z}^2 / \mathrm{var}\{z\} \quad . \tag{55}$$

"d^2" is an accurate and complete measure of performance if the correlator output z is gaussian, as will approximately be the case for large N.

The mean desired correlator output is given by

$$\bar{z} = \int_0^{N\Delta} a^2(t) \cos^2 \omega_0 t \, dt = \frac{N\Delta}{2} \quad . \tag{56}$$

Consequently,

$$d^2 = \frac{N}{(1 - \alpha)^2 + \alpha^2} \quad . \tag{57}$$

For $\alpha = 0, 1$, $d^2 = 4N$, and for $\alpha = 1/2$, $d^2 = 2N$. The mean performance measure is

$$d^2 = E_\alpha\{d^2(\alpha)\} = N \int_0^1 \frac{1}{(1 - \alpha)^2 + \alpha^2} d\alpha \quad . \tag{58}$$

Evaluating the intergral, one finds the mean or average performance specified by

$$\overline{d^2} = \pi N/2 \quad . \tag{59}$$

This performance is likely to be achieved on a steady-state basis when a large number of equal-power accesses are simultaneously present.* In this case, ensemble statistics will tend to produce the result of Eq. (59)--i.e., it will be very unlikely for all accesses to

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*More precisely, we mean that the undesired accesses have equal power and that the sum of these powers is equal to the power in the desired signal.

have simultaneous phase-chip transition or to have half-chip overlap between all accesses. Thus, Eq. (59) represents the fundamental multiple-access processing-gain capability of a binary, phase-reversal keying, phase-chipping, spread-spectrum system against similar signals. Table 5

Table 5

EFFECTIVE SNR FOR THREE CASES

Case	d^2
Best-case direct sequence	2N
Worst-case direct sequence	N
Average direct sequence	$(\pi/2)N$

presents the results for the three cases. The quantity d^2 is the effective SNR in the bit detector for the case of interference from a single undesired spread-spectrum access of power equal to the desired signal. It is interesting to note that the worst-case value is identical to the result of Eq. (11) of Aein.[3] Thus, on the average his results are pessimistic since they do not reflect the correlation spreading that occurs with nonsynchronous chip transitions.

Application of Results. The above results were derived for the case of a single equal-power interfering access. Thus the values of d^2 are also the values of the processing gain. In the prior simplified analysis the processing gain was denoted W/b where W and b were the RF and baseband bandwidths, respectively. At this point, we can say that the previous simplified result was correct for the worst-case analysis, provided W and b are defined properly.

Since the direct-sequence form of spread spectrum is assumed, both the RF signal spectrum and the matched-filter response will have a sinc-function-squared form. So long as the same definition of bandwidth for b and W is employed, the correct result is obtained. Thus, either equivalent-noise or first-null-to-first-null bandwidths may be used.

Consequently, the following basic expressions hold and may be used to estimate the probability of bit error based on the SNR d^2 at the correlator output:

$$d^2 = \begin{cases} P_S \Big/ \left[P_I\,(b/w) + bN_o\right] & \text{worst case} \\ P_S \Big/ \left[P_I\,(b/2W) + bN_o\right] & \text{best case} \\ P_S \Big/ \left[P_I\left(\dfrac{2b}{\pi W}\right) + bN_o\right] & \text{average case} \end{cases} \tag{60}$$

where b and W are both measured either as equivalent noise or as first-null-to-first-null bandwidths. Use of the worst-case formula represents a conservative approach. If one selects either of the other equations one must recognize that periodically the performance will degrade, perhaps by as much as 3 dB. The duration of these fades may be many thousands of data bits. Consequently, the short-term error rate can fluctuate significantly. As noted earlier for the case of a large number of users, the average-case formula represents the best choice since it will be very unlikely to find a significant fraction of the users similarly timed.

In conclusion, it should be noted that the above results were derived for unfiltered direct-sequence operation. For modest filtering (no tighter than the first nulls) the results will be approximately correct. However, with heavier filtering the chip transitions

become sufficiently "blurred" that the performance is essentially independent of the relative timing of chip transitions.

3. Power Control and Reliability

The power-control problem for SSMA is very similar to the problem for FDMA, since both techniques involve the simultaneous presence of angle-modulated signals in a transponder nonlinearity. The major difference stems from the significance of the access-noise (interference from other users of the system) effect with SSMA and FDMA. The fact that SSMA is a quasi-orthogonal approach implies that for the case of large dynamic range in signal levels, the power-control algorithm must reflect the access noise. This effect has been previously discussed in Section III-A-2-b. In contrast, it may be shown (see Section III-C-2-a) that access noise effects in FDMA may be made negligible (for power control) by the proper choice of transmit and receiver filters. Consequently, for FDMA the power-control algorithm need not reflect access noise.

Thus, the SSMA power-control approach consists of the FDMA power-control approach but modified to account for the access noise of SSMA. The reader is referred to Section III-C-3 for a discussion of FDMA power control.

So far, the discussion has neglected the problem of sensing the need for power-level adjustment. Section III-B-4-b presents a comparative evaluation of FDMA and SSMA with TDMA, with respect to power-control sensing.

The robustness of the power-control system is of considerable interest for a military environment. That is, how well does SSMA perform when there is unintentional error in the power settings of one or more accesses? It is difficult to give a precise answer without a more detailed description of the operating environment. Nevertheless, it is

possible to say that in general an error will not result in a catastropic system failure.

As an example of such problems, if there are many roughly equal power users and one transmits 6 dB too much power, then some power will be robbed from the other accesses and it will be necessary to reduce their data rate slightly. As another example, if there is one very strong access and it radiates 6 dB too much power, then it will be necessary to greatly reduce (by a factor of more than 4) the data rates of the other accesses. Nevertheless, communication remains possible at a lowered rate. Thus, one can see that unless there are very large power-control errors, some form of communication remains possible. Consequently, the network control system is not particularly vulnerable to catastropic failure. These same arguments will apply to FDMA as well.

4. Data-Rate Limitations

For spread-spectrum multiple access (SSMA) the data (bit) rate is normally much lower than the code (chip) rate. However, under some circumstances it may be desirable to raise the data rate to as high a number as possible. Thus, one is concerned with the limitation imposed by the SSMA format.

Since the code chips and data bits are mod-2 added, a masking effect results. If the code and data rates are equal, then each bit is perfectly randomized by one chip, and vice versa. Thus, it is impossible for the code-tracking circuits to establish or maintain lock.* Consequently, it is impossible for a SSMA modem to have a data rate equal to

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*The code tracking loops are of the delay-lock type using envelope correlators to account for the random data modulation. This circuit can be implemented either at baseband or at IF. The code rate must be higher than the data rate for this circuit to function. Its operation is predicated on a spectral-width reduction for properly timed codes.

the chip rate. However, in theory it is possible to operate with a data rate only slightly lower. This theory does not reflect various practical factors such as oscillator stability, acquisition time, and filter time constants. When these are considered, the minimum chip/bit ratio becomes 10. More typically the operating system values are in excess of 100.

By way of example, the highest data rate one might expect to achieve with a 40-MHz SSMA modem is 4 Mbps. If a higher bit rate is required, then multiple modems are necessary.

5. Anti-Jam Capability

SSMA is well known to possess inherent AJ capability, and extensive work has been done determining the vulnerability of various forms of SSMA to various forms of jamming. It is not our intent here to reproduce that work. Rather, the goal is simply to present some of the fundamental limitations as they apply to satellite communications. In the first subsection below, the significance of access noise on the interference immunity is evaluated. In the second subsection, the impact of a power-sharing nonprocessing transponder on the overall SSMA system jamming immunity is assessed. It is shown that SSMA does not yield the desired voice-channel capability for each access.

a. Effect of Access Noise on Interference Immunity

The access noise present from other quasi-orthogonal spread-spectrum signals does degrade the interference immunity of spread spectrum with respect to that available if truly orthogonal spread spectrum were used. Consider the case of an "ideal" linear AGC transponder. In this case the correlator output SNR is given by

$$SNR = \left[\frac{P_s}{(P_o + P_I + P_s)}\right] P_T \Bigg/ \left\{\left[\frac{P_o P_T}{(P_o + P_I + P_s)} + \frac{P_I P_T}{(P_o + P_I + P_s)}\right](b/W) + P_n\right\} \tag{61}$$

where P_s, P_o, and P_I use the powers at the transponder input of the desired access, other accesses, and interference, respectively; P_T is the total transponder power referred to the ground; P_n is the thermal noise power in the correlator bandwidth; and W/b is the processing gain. It should be noted that the harmful effect of access noise appears as the first term within the brackets of the denominator of Eq. (61). The presence of this term will cause the SNR to be lowered.

One can pose another way of viewing the impact of access noise. One can compare the allowable interference powers P_{I1} and P_{I2} for the case of orthogonal and quasi-orthogonal accesses, respectively, to produce a required SNR. The value of P_{I2} is smaller than P_{I1} and is given by

$$P_{I2} = P_{I1} - P_o \Bigg/ \left[1 + \frac{P_n}{P_T}\frac{W}{b}\right] \quad . \tag{62}$$

If the total satellite power referenced to the ground when reduced by the processing gain is small in comparison to the noise power, then there is little difference in tolerable interference levels.

b. Spread-Spectrum Performance with a Power-Sharing Satellite Transponder

The purpose of this subsection is to identify one of the fundamental limitations of SSMA modems operating with conventional, nonprocessing satellite transponders. Typically, these transponders

are driven into a nonlinear mode by the accessing signals.* In this mode of operation three effects take place: (1) creation of intermodulation cross-products, (2) suppression of weaker signals by stronger signals, and (3) power-sharing of the transponder power based on the ratio of desired signal power to the total power present at the transponder input. With SSMA modems the intermodulation products are not particularly significant, so the first item can be ignored in a first-order approximate analysis. The maximum suppression effect loss is 6 dB. This is certainly a significant effect, but it is bounded to be no more than 6 dB for any level or type of interference. By contrast, the power-sharing loss is unbounded and continues to increase with increasing interference power. Consequently, we concentrate our analysis on the limitation implied by this fundamental effect.

As the up-link interference power is increased, there are two effects. First, the additive interference power at the ground terminal is increased. However, the SSMA receiver has processing gain (roughly equal to the ratio of the RF bandwidth to the data rate) to combat this additive effect. With sufficient RF bandwidth, it is possible to obtain enough processing gain to render the effect of additive interference negligible.

The second effect is much more serious due to its multiplicative nature. As the interference power increases, the desired signal power decreases due to power sharing. The thermal-noise spectral density (N_o) at the receiver remains fixed, of course. Consequently,

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*Many transponders that are called linear are not, in fact. The most common example is the "linear" ideal AGC transponder where the gain is increased so that the full output power is achieved on an average basis. This mode of operation causes a power-sharing of the transponder's finite power resource that would not occur with a truly linear transponder. Thus, the ideal AGC transponder is nonlinear, since multiplicative effects take place between different accesses.

the down-link capacity quotient (C/kT or P_s/N_o) decreases with interference power. Shannon's theorem states that the achievable bit rate decreases in proportion to the capacity quotient. This result is independent of the modulation employed. Thus, it applies to spread-spectrum modems. Spread spectrum offers processing gain with respect to band-limited interference, but offers no advantage against thermal noise since it is not band-limited.* Consequently, strong up-link interference can cause the down-link to be thermal-noise-limited rather than processing-gain-limited. This problem is particularly severe for terminals that have low G/T.

The major conclusion is that spread-spectrum modems may not be able to take advantage of their full processing gain when used with satellite links that include a power-sharing effect due to the finite resources of the transponder. Consequently, when maximum interference is encountered for typical links, the maximum data rate is reduced to 75 baud rather than the 2400 to 9600 baud desired for a voice-channel capability.

At present this situation can be improved by: (1) increasing the satellite ERP (through either antenna gain or power), (2) increasing the ground terminal G/T, or (3) changing the character of the transponder. The first two items have obvious technology limitations and cannot be readily accomplished. If the satellite transponder is to

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*This may be seen in two ways. First, if the spread-spectrum-modem RF bandwidth is increased to obtain more processing gain, then more processing gain (in proportion to W) will be required since the RF thermal noise power increases in proportion to W. Second, it is well known that for antipodal binary signaling the error rate depends only on the ratio of energy per bit to noise spectral density and is independent of the signal waveform. Thus, the use of spread spectrum--i.e., a more complex, rapidly changing waveform--cannot improve the error rate as compared to simple binary phase-shift keying.

avoid significant power-sharing loss, a spread-spectrum receiver must be included at the transponder input so that processing gain can be used to greatly reduce the interference power entering the power-sharing portion of the transponder. Such a processing transponder has been suggested many times. It has not become an operational reality for the following main reasons: (1) equipment complexity, and (2) operational difficulties associated with all users employing the same code, or the requirement for multiple processors in the satellite.

Thus, one sees that SSMA can be used to provide a degree of interference immunity. This capability is not as great as one would like, due to the power-sharing effect in the satellite transponder. Nominally, one is only able to obtain 75 baud for the case of maximum interference. If this performance satisfies a requirement, then SSMA may be worth the price in equipment complexity, cost, reliability, and operational difficulty. However, if the true goal is to provide a voice-channel capability to each user, a high price is paid for inadequate performance.

B. Time-Division Multiple Access

In the following subsection, time-division multiple access (TDMA) systems are categorized as being one of two types. The second subsection discusses the timing accuracy achievable with a TDMA system, and the third analyzes the performance of identical and differing user TDMA systems and establishes the negligible effect of TDMA access noise. The fourth subsection studies the power-control problem of TDMA for different network timing systems and compares the power-control-sensing problem with FDMA and SSMA. The data-rate limitations of TDMA systems are briefly considered in the fifth subsection, and the AJ capability of TDMA is analyzed and mode-switching problems are considered in the sixth subsection.

1. Types of TDMA Systems

There are two fundamental types of TDMA systems. The categorization is based on the mode of operation of the carrier-phase-reconstruction loop.* With long-frame-duration and long-time-slot-duration systems it is possible to allocate a small portion of each time slot to acquisition of carrier phase and bit timing. When the next frame occurs the receiver must reacquire the carrier phase in his receive time slot. Thus, this is known as the reacquisition mode of operation. The advantage of the reacquisition mode is that each user can demodulate all signals with just one receiver. The disadvantage is that to obtain the desired efficiency of 90 to 95 percent with a reasonable number of accesses it is necessary to use quite long frames. This implies large and costly buffers and increased message delay.

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*Normally the bit-synchronization loop will be of the same type as the carrier-reconstruction loop.

For short frames and time-slot durations one cannot afford to provide sufficient time to reacquire within a time slot. It is necessary to maintain phase lock from one frame to the next. This is known as the inertial mode of operation. Figure 6 illustrates the timing format for these two fundamental approaches. Since each transmission will have independent carrier phase with respect to the others, it is necessary to have separate carrier-reconstruction loops for each signal that a user desires to demodulate. The advantage of the inertial mode of operation is the high frame efficiency (i.e., small wasted portion of the frame

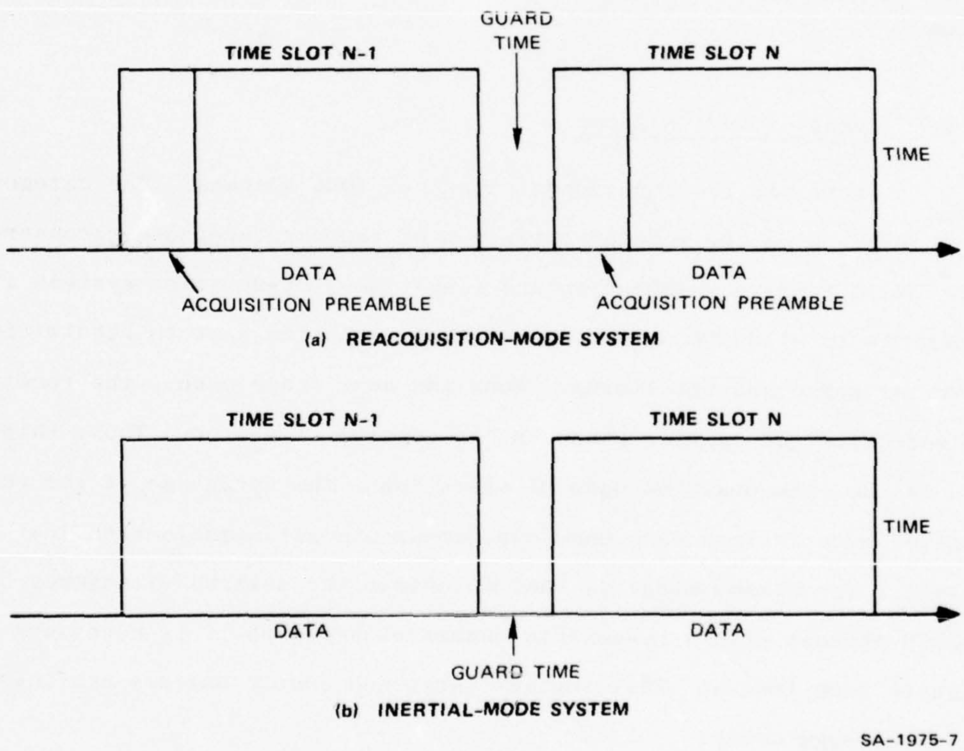


FIGURE 6 PORTION OF FRAME TIME ILLUSTRATING THE FORMAT FOR THE REACQUISITION-MODE SYSTEM AND THE INERTIAL-MODE SYSTEM

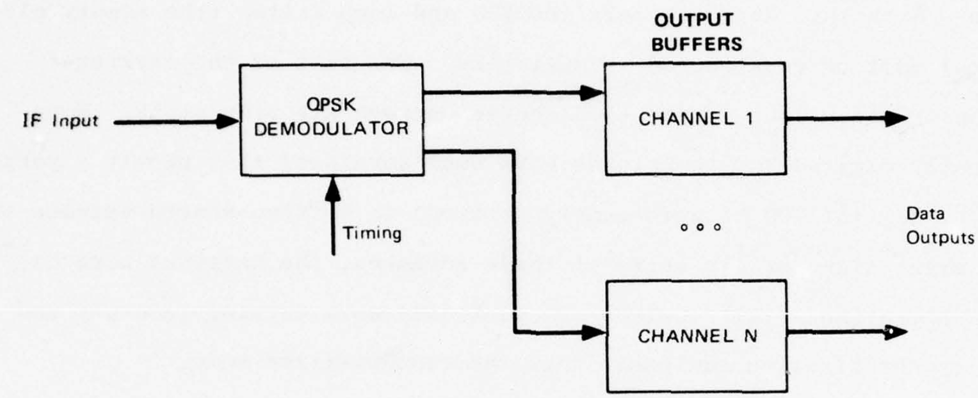
duration. The disadvantage is the requirement for multiple carrier-reconstruction loops to receive multiple transmissions. Figure 7 presents block diagrams for both reacquisition-mode and inertial-mode receiving systems. Note the greater complexity of the latter system. This problem has been improved somewhat by the development of multiplexed loop. With this approach only the VCO and loop filter (the memory elements) must be constructed in multiples. The rest of the carrier-reconstruction loop can be time-shared between all time slots. More recently digital VCO techniques have been developed that permit a portion of the digital VCO (a zero-memory portion) to be time-shared between the different signals. In spite of these advances, the inertial mode of TDMA operation still requires substantially more carrier-recovery and bit-synchronization equipment than the reacquisition mode.

2. TDMA Timing Accuracy and Reliability

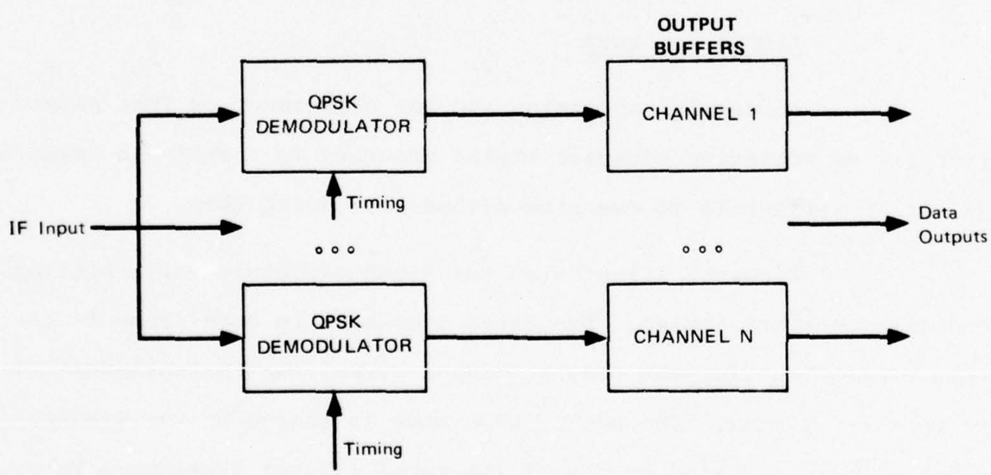
a. Timing Accuracy

Successful operation and the efficiency of TDMA depend critically on achieving adequate timing accuracy in a reliable fashion. Thus, it is worthwhile to describe methods for doing this.

Figure 8 illustrates one frame structure for achieving the desired network timing. The first time slot in each frame is reserved exclusively for the central timing signal that establishes network receiver timing. The second time slot is shared by the timing/ranging signals radiated by all of the local (slave) stations. These signals are required to establish the transmit-time base of the local stations. The rest of the frame is available for time-sharing among the user communication signals. The basic concepts[1,4] of the network timing system are illustrated in Figure 9. Consider the acquisition process at a local station. The central timing signal, a pulsed spread-spectrum



(a) REACQUISITION-MODE RECEIVER



(b) INERTIAL-MODE RECEIVER

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FIGURE 7 BLOCK DIAGRAMS OF THE REACQUISITION-MODE RECEIVER AND THE INERTIAL-MODE RECEIVER

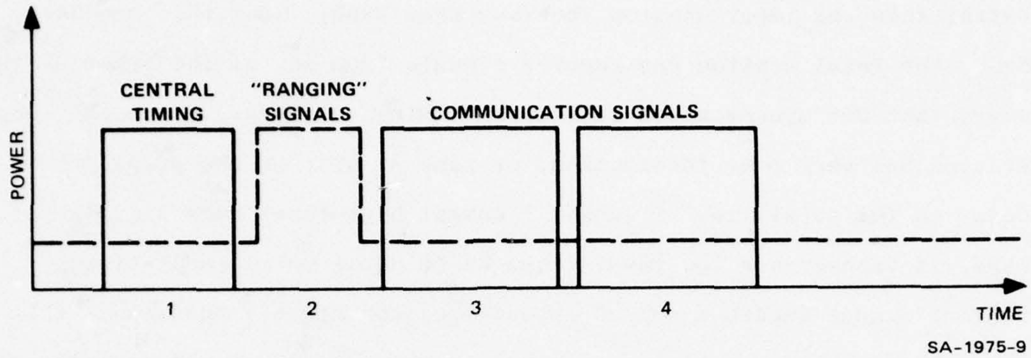

FIGURE 8 TIME-DOMAIN REPRESENTATION OF THE PROPOSED TDMA TIMING SYSTEM

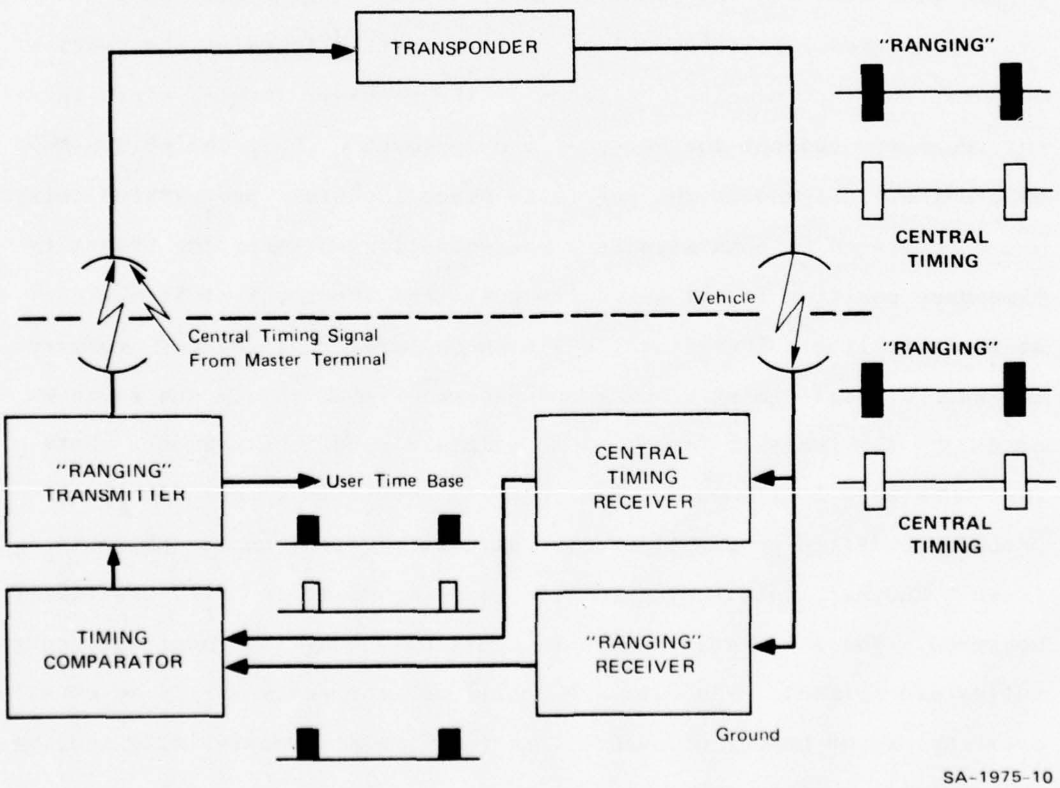

FIGURE 9 CONCEPTUAL BLOCK DIAGRAM OF THE NETWORK-TIMING SYSTEM

signal, is received by the local-station central-timing receiver. A narrow and precise pulse once per frame is derived in this receiver and establishes the local-station receiver time base. Once this has been done, the local station can receive signals from any of the other system users that are synchronized to network timing. However, since the local station has very poor information, or none at all, on the propagation delay to the satellite, it cannot transmit high-level data signals yet. Thus, it transmits a low-level (down by 10 dB or more) local-timing/ranging signal (again a pulsed spread-spectrum signal) and sweeps this transmit time base until its received timing/ranging signal is precisely aligned in the local-timing/ranging time slot. This, in effect, means that the narrow pulses corresponding to the received central-timing-signal time base and the received local-timing/ranging time base are exactly aligned. If these pulses are precisely aligned at the receiver outputs, they are precisely aligned at the receiver inputs, since identical receivers (except for PN code) are employed. Thus, the pulses must be precisely aligned at the satellite since identical propagation delays are encountered by both signals. Consequently, whatever the transmit-time-base position is, it correctly positions the local station signals at the satellite. Thereafter, a timing comparator, in effect, compares the narrow local-timing/ranging and central-timing pulses and advances or delays the transmit time base to maintain precise alignment. Note that this is a long-loop feedback-control system that includes the 1/4-second round-trip propagation delay as part of its transfer characteristic. Further, note that satellite range or delay is never explicitly measured. The local station can transmit data once the local and central pulses are aligned. The data are positioned correctly simply by counting over the proper number of quanta from the transmit local-timing/ranging narrow-pulse-marking frame start.

The timing accuracy of an implementation of this concept has been measured.[5] These measurements were performed with IF simulation tests. Consequently, the effects of dynamic peak variation were not assessed. However, the loop bandwidths are sufficiently great that the transient errors should not be significant. No difficulties were encountered with this experimental system.*

The timing system performs remarkably accurately. Figure 10 lists the measured peak-to-peak short-term jitter (in nanoseconds) of the central timing receiver as a function of frame rate and capacity quotient. The long-term jitter was assessed on an overnight basis at 600 frames per second (fps) and a capacity quotient of 74 dB (Hz).

C/kT dB (Hz) \ FRAME RATE	600	1200	2400	4800	9600
70	8	6	4	2	0.5(+)
80	4	3	1	1(+)	0.5(-)
90	1	0.5	0.5(-)	1	0

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FIGURE 10 PEAK-TO-PEAK JITTER (IN NANOSECONDS)

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*With this system the timing accuracy is dependent on the frame rate, since the timing-signal code rate was a fixed factor greater than the frame rate. With a fixed-frame-rate, nonexperimental system, the code rate and the timing accuracy could be made equal to the best values measured.

For this test, the peak-to-peak jitter was found to be 20 ns, or 2.5 percent of a timing-code chip.

Timing accuracy of the local transmit time base with respect to the local receive time base was also measured experimentally. Figure 11 shows the measured results in the IF tests as a function of capacity quotient and frame rate. The ordinate axis is the standard deviation of the timing error in nanoseconds. These tests were performed (except for the one case so noted) with the local-timing signal suppressed by 10 dB with respect to the data signal and the capacity-quotient measurement. Since there is no limiter in the IF channel, these tests are pessimistic by 10 dB. These results show highly accurate network timing with

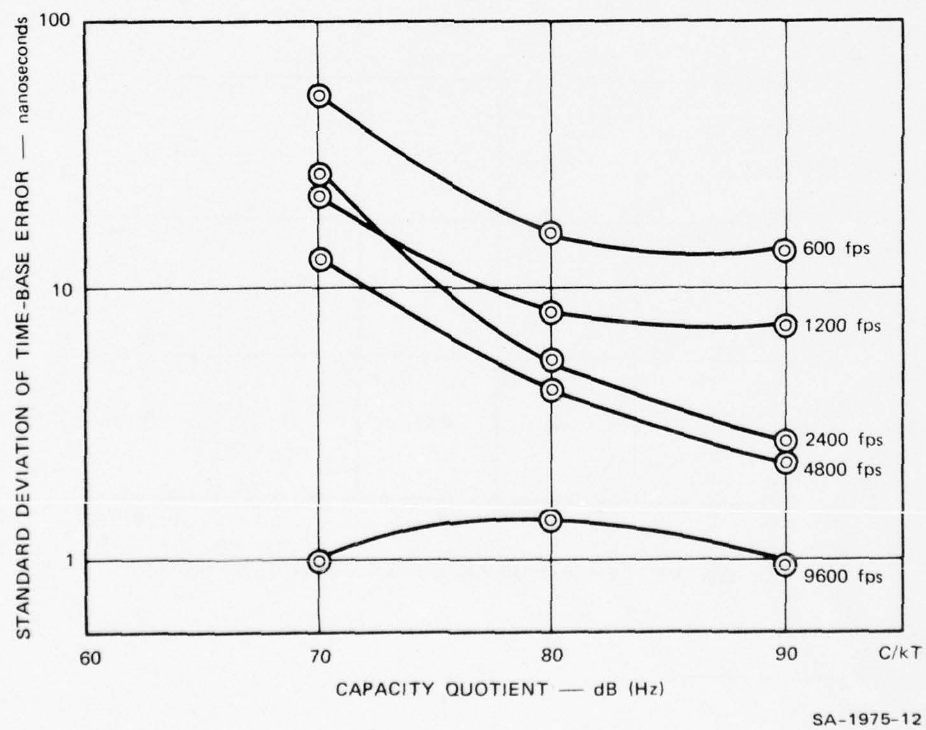

FIGURE 11 STANDARD DEVIATION OF TIME-BASE ERROR AS A FUNCTION OF CAPACITY QUOTIENT AND FRAME RATE FOR LOCAL-TRANSMITTER TIME BASE AND LOCAL-RECEIVER TIME BASE

the 1-ns performance at 9600 fps being particularly impressive. At 600 fps the timing accuracy degrades, due to the lower code rate, to a standard deviation of 13 ns for an asymptotic value at high-capacity quotients.

Similar excellent results are reported by Sekimoto and Puente of COMSAT Corp. using a network-timing-system concept based on a unique word detection with the data stream.[6] They report the possibility of guard times of 100 ns between bursts with synchronous-altitude satellites. An identical conclusion is reached by Nosaka et al.[7]

Consequently, it may be stated that achieving the desired timing accuracy presents no problem for TDMA systems. The presence of large differences on power levels between adjacent channels may make a larger guard time desirable to avoid transient decays. Calculations presented in another section indicate that a guard time of 500 ns is more than adequate.

b. Timing-System Reliability and Robustness

It should be noted that the network power-control system of FDMA and SSMA and the network-timing-control system of TDMA play analogous roles. It has been previously noted that the network-power-control system is not particularly vulnerable to catastrophic failure, but in a sense degrades somewhat gracefully. By contrast, the network-timing-control system is potentially capable of causing catastrophic failure of the TDMA system. In fact, the network-timing system can be

correctly viewed as being an all-or-nothing system.* As a result, designers of TDMA systems have given serious consideration to the reliability and robustness of the network-timing system.

For example, early experimental TDMA systems included the following features. First, since central timing is a required function (the most crucial) for every user (even receivers only), it was allotted an additional margin of power. Furthermore, TDMA modems were designed so that any terminal could operate as the central-timing station in case of failure of the primary central-timing station. Second, both the central-timing and local-timing subsystems were designed so that their thresholds were more than 10 dB below that of the TDMA data signals. This was possible due to the low tracking bandwidths required for the timing signals. Third, spread-spectrum formats were employed on both of the timing signals to enhance their interference immunity. Fourth, a failsafe mechanism was employed such that up-link transmission could not commence unless: (1) central-timing-signal lock was achieved; (2) local-timing/ranging-signal lock was achieved, (3) the local operator desired to transmit, and (4) the central control station gave permission for transmission. The last item provides a means for controlling users when their local sensors have failed and proper timing is erroneously indicated.

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*The situation is actually not quite this black and white under all circumstances. For example, with very high burst rates the network-timing accuracy may be sufficiently good to prevent pulse overlaps but inadequate to unambiguously (with low probability of error) identify the first data bit in a time slot. In this case a time-base error might result, yielding a 50-percent bit error rate for that burst. Practical TDMA systems incorporate special subsystems for making the probability of this event (sometimes called bit count integrity) significantly lower than the probability of a bit error. Thus, the threshold of catastrophic failure due to this phenomenon is placed far below the data-system threshold.

Thus, we see that by proper design it is possible to make the network-timing system more robust than the TDMA data system. This is necessary in any military TDMA system due to the catastrophic-failure mode of network timing.

3. Multiple-Access Capability

The first subsection below analyzes the throughput achievable with a TDMA system based on identical accesses. The second subsection compares the performance of TDMA and FDMA when there are differences between the G/T at the receiving stations. It is assumed that all accesses desire the same rate. It is shown that the performance of FDMA and TDMA are essentially identical for the anticipated operating situation. The third subsection analyzes the significance of access noise (interference from previous time slot) for TDMA. The analyses of this section are closely related to the power-control problem for TDMA, which is discussed in Section III-B-4.

a. Analysis of TDMA-System Performance for Identical Accesses

Time-Division Losses--The analysis of the multiple-access capability of a PSK/TDMA modulation technique is quite simple. The communication performance is determined by the effective down-link capacity quotient, $\hat{Q}_d$, rather than the down-link capacity quotient, Q_d. These quantities are related by the expression

$$\hat{Q}_d = L \cdot Q_d \cdot \beta \qquad (63)$$

where L and β are loss factors described below.

The loss factor, L, may be usefully considered to be a product of three elemental loss factors:

$$L = L_1 \cdot L_2 \cdot L_3 \quad . \tag{64}$$

The elemental loss factors are

$$L_1 = 1 - \left[(N + 2)\,\frac{t_g}{T_f}\right] \tag{65}$$

where N is the number of useful access channels, t_g is the required channel guard time due to path delay uncertainties,* and T_f is the frame period. The factor (N + 2) rather than N arises since it is assumed that two control channels--i.e., two time slots--are required. Thus, L_1 corresponds to the guard-time loss.

The control-channel's loss is given by

$$L_2 = \frac{N}{N + 2} \quad . \tag{66}$$

It is to be noted that Eq. (66) is meaningful only for N much larger than 2. For example, it would be foolish to employ two time slots to control one other slot. For the values of N of interest, Eq. (66) is reasonable.

The control-channel loss is included with some hesitation, since analyses of other multiple-access systems may not include this loss. For similar operating conditions consisting of very-high-duty-factor usage, all techniques require control channels. For example, in a high-usage-factor environment, the addition of one more SSMA signal may seriously degrade the whole system. Thus, it is necessary to obtain permission via a control channel to transmit. Since control channels appear advisable, the control-channel loss is included here:

$$L_3 = \alpha \left[1 + \frac{W_s}{Q_u}\right]^{-1} \quad . \tag{67}$$

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*If the TDMA system uses a preamble, this wasted time should be included in the value t_g.

L_3 is the power-sharing-and-suppression loss factor. Q_u is the single-user up-link capacity quotient. W_s is the satellite bandwidth, and α is the desired-signal-suppression factor, ranging from 1.0 to 0.8, under normal circumstances. Typically, $W_s \leq 0.1\ Q_u$. Consequently, the power-sharing loss factor, L_3, is usually negligible.

The other loss factor, β, is due to interfering noise power transmitted by the satellite transponder, and is given by

$$\beta = \left[1 + \frac{N_o^1}{N_o}\right]^{-1} \tag{68}$$

where N_o^1 is the received noise-power spectral density from the satellite transponder. The ratio of noise-power spectral densities is given by

$$\frac{N_o^1}{N_o} = \frac{\gamma Q_d}{W_s}\left[1 + \frac{Q_u}{W_s}\right]^{-1} = \frac{\gamma Q_d}{W_s + Q_u} \tag{69}$$

where γ is the noise-power-spectral-density suppression factor, ranging from 1.0 to 0.5 under normal circumstances. If $Q_u \geq 10\ W_s$, as normally would be the case, then $(N_o^1/N_o) \approx \gamma\ Q_d/Q_u$. Again, typically, a down-link capacity quotient is 20 dB inferior to an up-link capacity quotient. Thus, normally the loss factor, β, is also negligible.

Consequently, under normal operating conditions, the major losses are the guard-time loss and the control-channel loss. The latter is given directly by Eq. (66). However, the former loss is determined only after the ratio of guard time to frame time is specified. Typically, this ratio ranges between $4 \cdot 10^{-4}$ to $1 \cdot 10^{-3}$. For example, a guard time of 0.5 μs and a frame time of 417.7 μs yields $1.2 \cdot 10^{-3}$. Thus, for a moderate number of accesses, the guard-time loss is negligible. Consider 47 access channels, $t_g = 0.5$ μs, and $T_f = 417$ μs; then $L_1 = 0.94$ or -0.25 dB.

For these parameters, the control-channel loss is given by $L_2 = 0.96$, or -0.17 dB.

Thus, for the example cited, the total loss factor is less than 0.5 dB for typical Q_d and Q_u. Consequently, it is seen that TDMA can be a very efficient multiple-access technique.

It is to be noted that no loss has been associated with the central timing signal or frame beacon pulse. This is permissible, since it is possible to utilize the satellite beacon signal for timing signals. That is, the satellite beacon signal power is required (or at least desired) for angle-tracking purposes. By suitable modulation, all or an appropriate portion of the beacon signal power may be made available for frame-timing purposes.

Transponder Data Rate--Given Q_u, Q_d, N, t_g, T_f, and W_s, the effective down-link capacity quotient, $\hat{Q}_d$, may be determined by Eqs. (63) through (69).

The total effective transponder data rate is given by (assuming no satellite bandwidth limitation and identical accesses)*

$$R = N \cdot r = \frac{\hat{Q}_d}{(E/N_o)^*} \tag{70}$$

where r is the user data rate and $(E/N_o)^*$ is the minimum energy-to-noise-power spectral-density ratio (one-sided) to achieve the desired bit error probability.

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*Identical accesses are assumed for simplicity. The analysis could be extended readily to cover unequal data rates and capacity quotients.

For a bit error probability of 10^{-5}, phase-shift keying (binary or quaternary) requires $(E/N_o)^* = 9.6$ dB. Thus, the total effective transponder data rate is approximately 10 dB less than the effective down-link capacity quotient. The individual data rate is degraded by a factor of N.

Sufficient information--i.e., Eqs. (63) through (70)--has been given to determine R and r versus N. In order to aid the reader in gaining intuitive insight into TDMA performance, a plot of R and r versus N, in a normalized form, is presented in Figure 12. The abscissa is the number of access channels, N, and the ordinates are the normalized data rates, $R(E/N_o)^*Q_d$ and $r(E/N_o)^*Q_d$. Thus, the entire loss is given by L_1L_2.

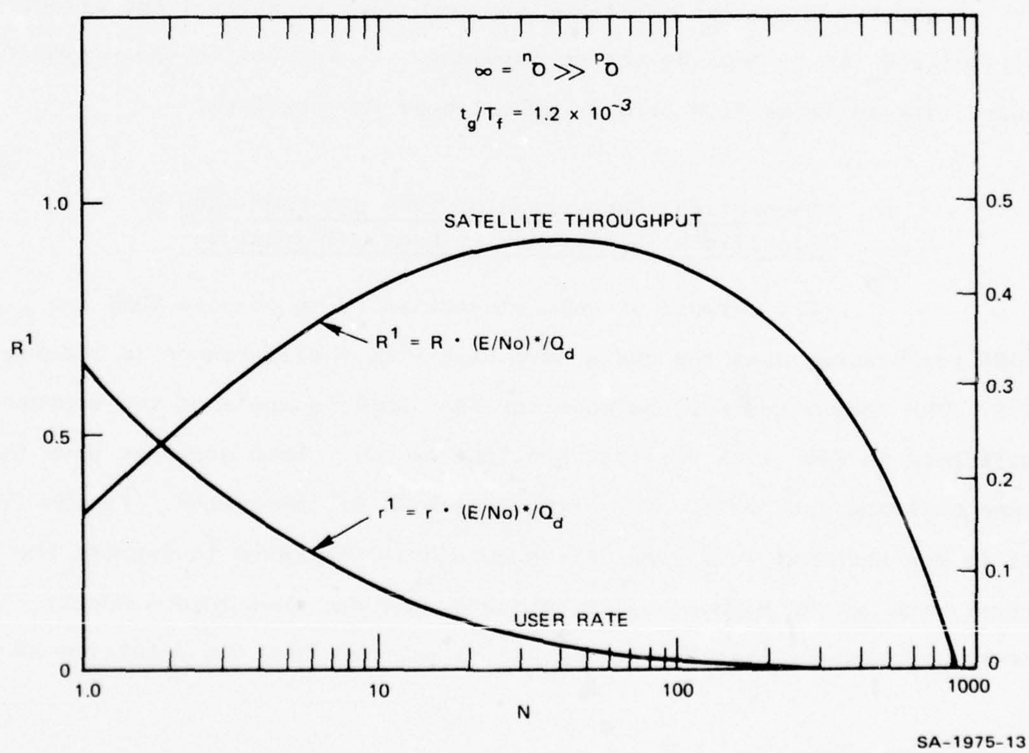

FIGURE 12 NORMALIZED DATA RATES VERSUS NUMBER OF IDENTICAL ACCESS CHANNELS

Again, it is to be noted the curve of R is misleading in the region of small N. That is, it is not necessary to use such large time slots for the two control channels as those used for the single (N = 1 case) communication channel. Note that the number of access channels may often markedly exceed the number of accesses. This is the case because one high-data-rate access may require many access channels--i.e., time slots. Consequently, the plot of R is reasonable in the region of interest. Note the existence of an optimum (in the sense of maximum transponder data rate) number of access channels. The optimum occurs near N = 50 but is quite broad.

A vast number of graphs similar to Figure 12 could be generated for different parameter values. However, this would serve no useful purpose since the equations are very easy to apply. The effect of finite Q_u is to degrade the performance. Variations in the ratio of guard time to frame time stretch or collapse the abscissa.

b. Theoretical Comparison of TDMA and FDMA when a Significant Difference in User G/T's Exists

The purpose of this subsection is to compare TDMA and FDMA performance when the users have significant differences in receive G/T. The comparison will be made for the simple example of two accesses differing in down-link capacity quotient by ΔQ.* Both accesses have the same baseband data rate. Two situations will be considered. In the first, it is assumed that sufficient RF bandwidth is available to support the burst rate to the high-capacity quotient user and that lower-capacity users are present. In the second, it is assumed that the satellite ERP

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*This example is sufficient to clearly demonstrate the fundamental effects. The general case of n users is analyzed in Appendix B.

is sufficiently great that the major restriction on satellite throughput is the available RF bandwidth.

Consider a TDMA system for the first situation. The total frame duration T_F will divide into two parts, T_1 and T_2, for the high (Q_1) and low ($Q_2 = Q_1/\Delta Q$) capacity accesses, respectively. The baseband data rate of each of these accesses is desired to be the same, but the RF burst rates will be R_1 and R_2, where, in general, $R_1 \neq R_2$. Figure 13 illustrates this.

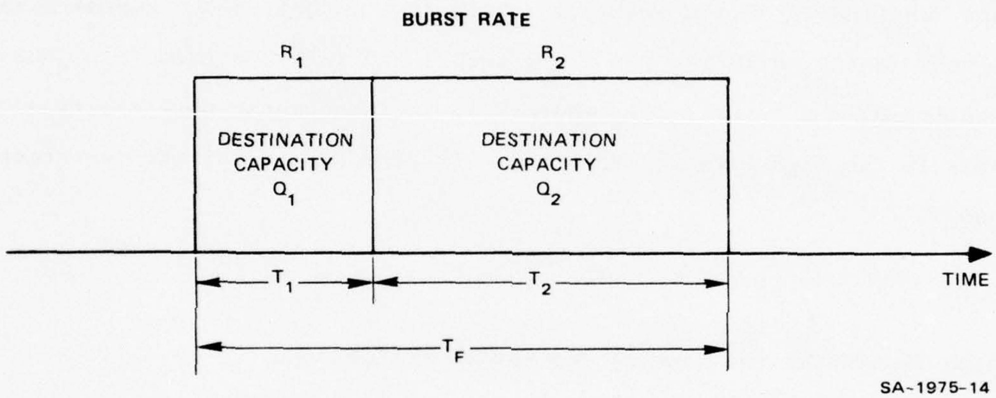

FIGURE 13 ILLUSTRATION OF TDMA FRAME FORMAT

The average satellite throughput is given by

$$TP = R_1 \cdot (T_1/T_F) + R_2 \cdot (T_2/T_F) \tag{71}$$

where $R_1 = R$ the maximum rate determined by the highest-capacity quotient, Q_1, and $R_2 = R/\Delta Q$. Since $T_1 + T_2 = T_F$, one finds that $(T_1/T_F) = 1/(1+\Delta Q)$ and $(T_2/T_F) = \Delta Q/(1+\Delta Q)$. Using these results, the satellite throughput is given by

$$TP = [2/(1+\Delta Q)]R \quad . \tag{72}$$

As an example of the harmful effect of G/T disparity, consider the case where ΔQ is 9.5 dB. In this case the throughput is reduced by a factor of 5 from that obtainable with the higher-capacity quotient link. Thus, the baseband data rates for both accesses are equal and are reduced by a factor of 5.

An analogous result obtains for FDMA. Here the proportion of power (rather than time) devoted to the weaker link must be increased by a factor proportional to the disparity in the capacity quotients. Thus, we find $(P_1/P_T) = 1/(1+\Delta Q)$ and $(P_2/P_T) = \Delta Q/(1+\Delta Q)$. Furthermore, we know that $R_i \sim P_i(G/T)_i$ for i = 1 and 2, and $(G/T)_1 = \Delta Q(G/T)_2$. Consequently, $R_1 = R_2 = R/(1+\Delta Q)$, where R is the throughput on a single-access basis to the high-capacity terminal. The throughput for the two-access case is

$$TP = [2/(1+\Delta Q)]R$$

which is exactly the same as for the TDMA technique.

Some have argued that TDMA is inefficient in bandwidth utilization since it does not transmit at the highest rate (occupy the full bandwidth) for the entire frame (see Figure 14). However, note that there is insufficient power to place an additional channel in the unused portion of the spectrum. In other words, low bandwidth utilization has little significance for a power-limited link. A similar problem occurs with FDMA. The lower-capacity quotient lowers the common bit rate of both accesses. However, with FDMA the occupied bandwidth is less than with TDMA.

Now consider the second situation, in which the transmission rate is limited by the available RF bandwidth. For the two-access TDMA systems there are several possible impacts for a disparity in capacity quotients. For example, if ΔQ is small enough, an adequate

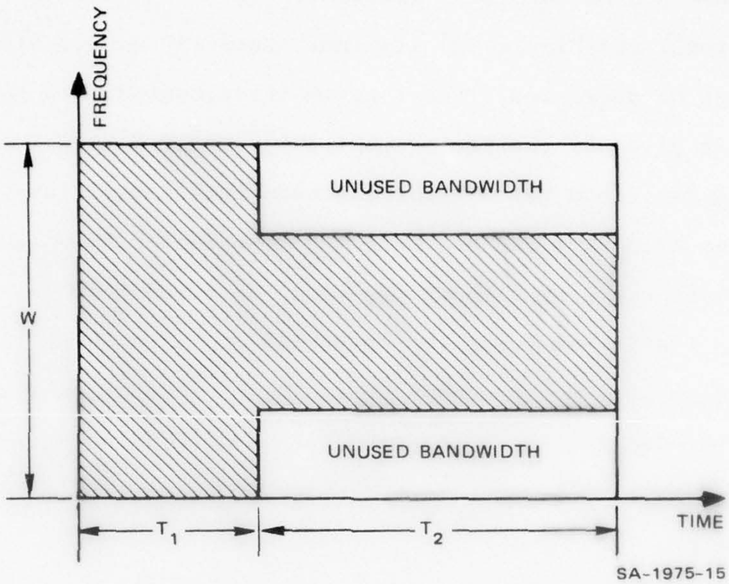

FIGURE 14 TIME-BANDWIDTH DIAGRAM ILLUSTRATING UNUSED BANDWIDTH DUE TO DIFFERENCE IN G/T BETWEEN RECEIVERS

E/N_o will be maintained to support the bandwidth-limited burst rate. Consequently, ΔQ will have no harmful effect on the TDMA system. However, if ΔQ is sufficiently large, it will be necessary to reduce the RF burst rate to this user. For this access, transmission is power-limited rather than bandwidth-limited. In this case there are several possibilities. First, one could proportionately lower the baseband data rate to this user. Second, one could simply eliminate this user, since it is impossible to provide the desired base-band rate. Third, one could distribute the frame time so that each user obtained the same baseband data rate. In this case, the weaker link would receive the greater portion of the frame time, during which it would transmit at a lower

burst rate. Since the second access is given a greater portion of the frame, the first access must receive a lesser portion. Consequently, since the burst rate of the first access cannot be increased due to bandwidth limitations (see Figure 14) its individual throughput--i.e., baseband data rate--must be decreased. The loss in throughput due to this mode of operation is given by the factor $[2/(1+B)]$, where B is a factor related to ΔQ as described below. For the bandwidth-limited case, ΔQ is composed of two factors A and B. Part A is the factor by which the capacity quotient can drop without producing an error rate in excess of that specified. B is the remaining factor in ΔQ. The burst rate for the weaker access must be reduced by this factor. Consequently, we see that for the bandwidth-limited case the TDMA system is not hurt as much, in general, by the capacity-quotient disparity, as for the power-limited case.

Now consider an FDMA system faced with the same situation. That is, the capacity-quotient disparity is sufficiently large that the weaker link cannot support the data rate possible with the available RF bandwidth. In this case one can do several things. First, one might simply lower R_2 to an acceptable value. Second, one might simply remove the second access. Third, one could redistribute the power so that the weaker link could, in fact, support the desired data rate. For the bandwidth-limited case there may well be sufficient link margin that this could be done without reducing the power in the stronger access to a point where it was unable to transmit at a desired rate. Figure 15 illustrates the spectra for this situation. For such circumstances, FDMA has an obvious advantage over TDMA, since it is possible to maintain the maximum throughput (as limited by the available bandwidth) in spite of the disparity in capacity quotients.

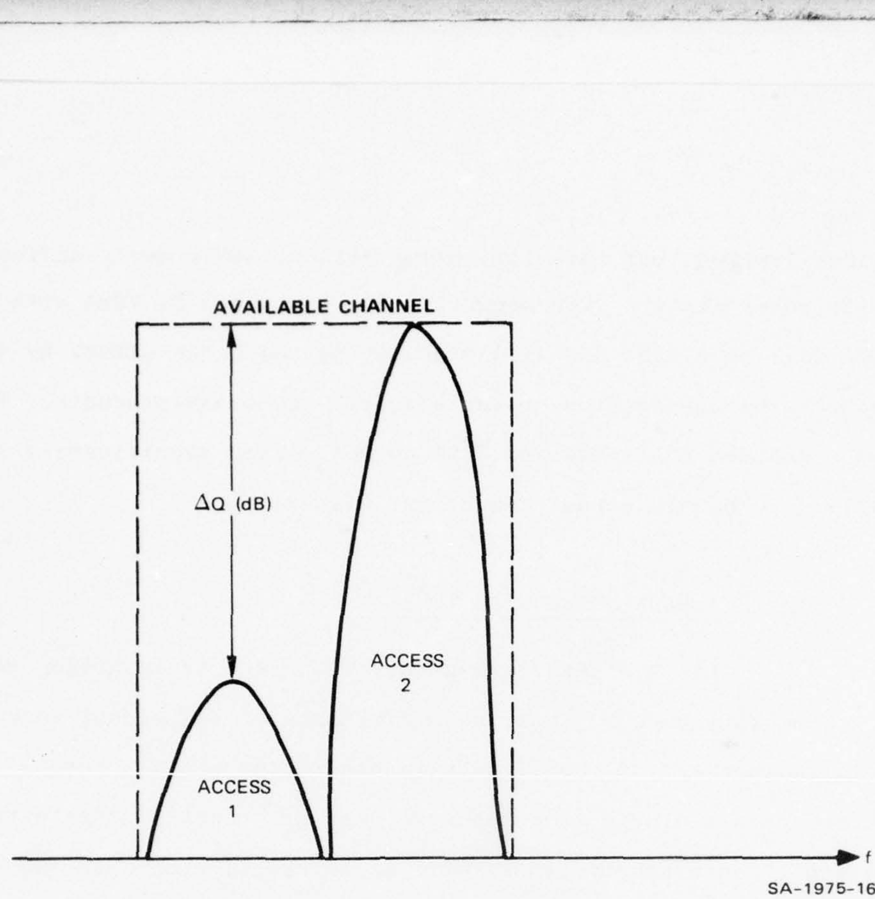

FIGURE 15 SPECTRAL DIAGRAM ILLUSTRATING HOW A BANDWIDTH-LIMITED FDMA SYSTEM CAN ACCOMMODATE A DIFFERENCE IN RECEIVER G/T's WITHOUT AFFECTING THE TOTAL THROUGHPUT

Thus, one sees that there are some circumstances under which FDMA is superior to TDMA with respect to throughput.* These conclusions are based on an idealized satellite transponder and do not include the effects of suppression, power sharing, and intermodulation cross-products. When such effects are considered, the relative advantage of FDMA will be reduced. Furthermore, the likelihood of a satellite system that is severely bandwidth-limited, as opposed to power-limited, must be considered. It is anticipated that future systems will be

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*The above argument assumes that the same modulation format is maintained for all situations. Theoretically, one might consider trading energy efficiency for bandwidth utilization by using different modulation--e.g., 4-ary or 16-ary PSK, for different accesses. This approach is not considered here for reasons of operational complexity.

bandwidth-limited, but this limitation will not be greatly different than the power limit.* Consequently, the advantage of FDMA with respect to TDMA will be slight and will probably be more than offset by the harmful effects of suppression, power sharing, and cross-products. Thus, it may be concluded that FDMA and TDMA do not differ significantly in their capability to handle capacity-quotient disparity.

c. Access Noise with TDMA

It is often claimed that TDMA is a truly orthogonal access system in the sense that the performance of the signal in one time slot is independent of the signals in other time slots. For infinite bandwidth this claim is true; however, all practical systems have finite bandwidth. Consequently, there will be interaction between the different users in the form of access noise.

The most significant access-noise effect results from the transient decay of energy from the previous time slot. A similar effect occurs for all time slots, but the decay for the non-adjacent slots is sufficiently great (in a reasonably performing system) that their effect is negligible. The major point is to determine if the guard time between adjacent slots is sufficiently large to permit the decaying energy to be negligible in comparison to the desired signal.

This transient-recovery analysis must reflect the potential power-level difference between adjacent accesses. The situation is obviously the worst when the weakest signal is preceded by the strongest

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*This is true for several reasons. First, as satellite ERP increases, there will be a tendency to reduce the receive antenna size to enhance terminal mobility. Second, as bandwidth limitations become more severe, higher-order modulation systems will be employed that are less power-efficient.

signal. Figure 16 illustrates this case. Fortunately, for the case of TDMA this power-level difference is likely to be considerably less than for FDMA. With FDMA the power level is adjusted to account for data rate and receiver sensitivity. By contrast with TDMA, these variables are accommodated by adjusting the time-slot duration rather than the peak power. Consequently, the up-link power-level variations in a TDMA system are due solely to: (1) imperfect control of the transmitter power output, (2) poor antenna pointing, and (3) propagation losses. These variations are significantly less than those dictated by data-rate and receiver-sensitivity variations. The former may be conservatively estimated to be no more than 10 dB. Thus, TDMA has a much less serious dynamic-range problem than FDMA and SSMA.*

Perhaps the best way to determine the significance of access noise in a TDMA system is to evaluate an example. Assume that a 10-dB power-level difference exists between adjacent time slots, with the more powerful signal preceding the weaker. Let the significant band-limiting be performed by the filter at the ground transmitter. This filter is assumed to have a 50-MHz, 3-dB bandwidth and to be of a 7-pole, 0.1-dB-ripple, Chebychev design. The transient decay will be considered to be negligible when it is 10 dB or more lower than the desired weaker signal power. The object is to determine the required guard time such that this goal is obtained.

The transient decay can be determined by using the published curves by Zverev.[8] For the above-described 7-pole Chebychev filter, the transient decay is less than 10 percent (power down by 20 dB) for normalized time in excess of 13 s (see Figure 17). The curve is

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*Under some circumstances, such as aircraft transmission, the up-link power may be very low. In such cases, TDMA may have a significant dynamic-range problem.

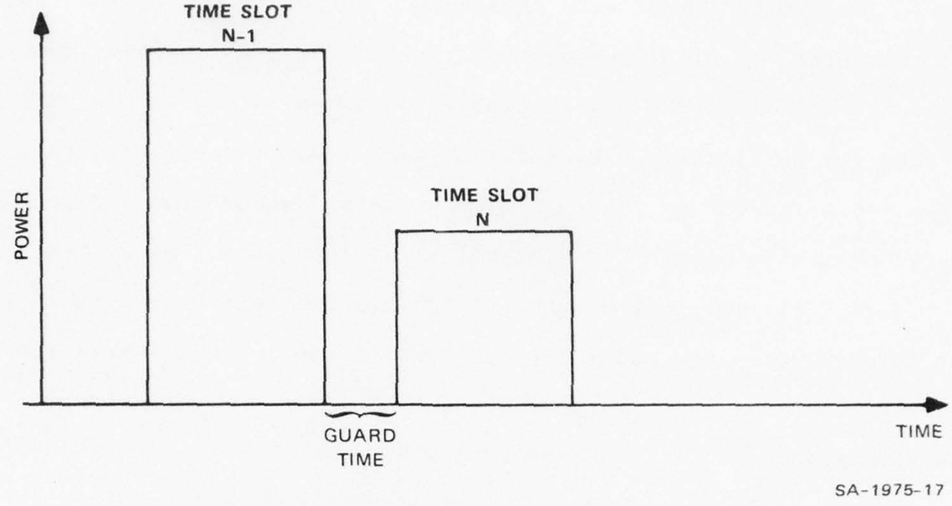

FIGURE 16 TIME-DOMAIN POWER DIAGRAM ILLUSTRATING THE WORST-CASE SITUATION FOR ACCESS NOISE DUE TO TRANSIENT DECAY

plotted for a single-sided 3-dB bandwidth of 1 rad/s. For a two-sided bandwidth of 50 MHz the decay *time* is 80 ns.

This decay-time requirement is sufficiently small that, in combination with the maximum timing uncertainty, it permits a guard time of 0.5 μs to be employed with negligible access-noise effects. Typically, the timing accuracy of a TDMA system should be better than 100 ns for the worst case. (See Section III-B-2 for a discussion of timing accuracy.) Consequently, it is possible to either tolerate larger power-level differences with a 0.5-μs guard time or to shorten the guard time to approximately 180 ns. Furthermore, the above estimate is quite conservative in that the effect of access noise on the preamble bits within the time slot is not as devastating as on the data bits. By the time the data bits commence, the transient energy has further decayed.

It should be noted that a 0.5-μs guard time permits 200 independent time slots with 90-percent efficiency in a TDMA system with a 1-ms frame time. Consequently, the guard time determined in the above example is not excessive.

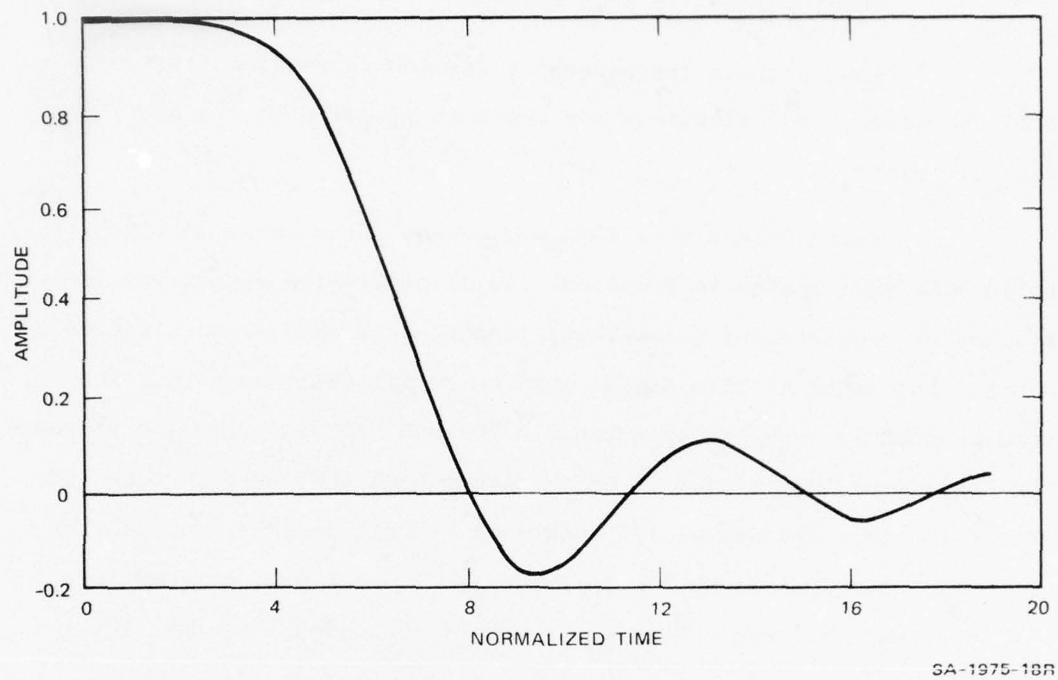

FIGURE 17 TRANSIENT DECAY OF A 0.1-dB-RIPPLE, 7-POLE CHEBYCHEV FILTER

In conclusion, the example evaluated above illustrates that most reasonable TDMA systems are essentially orthogonal and that access noise can be neglected.

4. Power Control

The first subsection below considers the fact that many forms of TDMA systems do require some form of power control in spite of the usual claim that no power control is required. In the second subsection, the power-control sensing or monitoring problems for SSMA and FDMA systems are compared with those of TDMA.

a. Power Control in TDMA Systems

TDMA systems are generally assumed to require little power control. However, two situations can occur in practice that greatly weaken this claim.

First, since most TDMA systems do not possess AJ capability, a hybrid TDMA/SSMA system is proposed. An AJ order-wire capability is maintained by transmitting a low-level (continuous) SSMA signal from each terminal. The level of this signal must be sufficiently high that the processing gain is adequate to support a 75-baud capability in the presence of the high-level TDMA signals. On the other hand the level must be sufficiently low that the sum of all such SSMA signals does not degrade the error rate of the TDMA signals significantly.* Under most circumstances such a level can be found. However, it should be noted that this level assumes perfect power control. The presence of power fluctuations greatly complicates the problem. With sufficient variation it may be impossible to find a level that will permit 75 baud for all cases without significantly affecting the TDMA error rate. Thus, power control may be required for this TDMA system.

A similar effect can take place with the network-timing signals. In fact, the above order wires have been frequently proposed as a method of establishing network timing. However, the same effect occurs

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*There is a natural human tendency to try to obtain performance within a few tenths of a decibel of theory. If one adopts this attitude, then the power level of order wire must be very, very low. However, if a more relaxed attitude is adopted, then the order-wire level can be increased substantially. For example, at an error rate of 10^{-5}, a 1-dB degradation will result if the total order-wire power is down by about 16 dB. Appendix C presents some sample calculations on the performance of the timing system in the presence of the TDMA-signal interference.

with more general network-timing signals. The problem is certain to occur with the local-timing signals and frequently occurs with the central-timing signal. Consider the local-timing signals. One proposal for reducing their effect on the data signals is to reserve a time slot for them exclusively. That is, a time-gated local-timing signal is used. Unfortunately, during initial acquisition, it is not usually possible to place the timing signal within the correct slot. Thus, the harmful interaction takes place; however, it occurs only during acquisition, which is, hopefully, infrequent and normally only involves one interfering signal at a time. Consequently, the interference level is reduced by a factor of N, the number of users.

Another local-timing-signal concept operates by tracking network timing on one's own data transmission.* This approach avoids the interference problem except during initial acquisition, which hopefully involves only one signal at a time. Nominally this will require transmission of a spread-spectrum signal. If the satellite ephemeris data can be provided to the ground terminal with sufficient accuracy with respect to the minimum time-slot size, then it is possible to completely avoid the use of a separate timing signal.† In this case, the power-control problem for TDMA is virtually nonexistent.

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*More precisely, the tracking can be performed on the bit-sync-clock preamble of one's own transmission.

†Due to orbit inclination effects, even the so called "stationary" satellites do exhibit substantial (may be nearly 2 ms) diurnal-delay variation. Fortunately, recent developments in station-keeping technology (TELESAT CANADA) promise substantial reductions in path-length variations. Consequently, the ephemeris data used must reflect the diurnal variations, which are continually changing in magnitude (see Appendix D).

There are other methods for avoiding power control for TDMA but they generally have some undesirable aspect. For example, power control can be avoided by using a separate transponder for the timing and/or spread-spectrum signals. The disadvantages are the requirement for a separate transponder and the need for additional bandwidth or the reduction of the bandwidth available for data transmission. The need for a separate transponder can be avoided by reserving a portion of the spectrum for the ranging signals and using the remainder for the data signals. The disadvantages of this approach are (1) for good TDMA timing, a large bandwidth is required, and (2) power-sharing and suppression still exist within the common transponder. The advantage is that the additive interference effects can be avoided by using separate frequency bands.

At present none of the above techniques for avoiding power-control interaction between data and network timing signals have proved practical or desirable enough. Consequently, one can anticipate that power control must be provided with TDMA systems. Future systems may avoid the problem, perhaps, by employing accurate ephemeris data and continuously "ranging" on one's own transmission.

b. <u>Power-Control Sensing for SSMA (CDMA), FDMA, and TDMA Systems</u>

The purpose of this subsection is to compare the difficulty in determining, at a central control station, whether adequate power control is obtained. Thus, we are concerned with the sensing problem and the effect of the multiple-access technique on this problem.

First, we observe that with TDMA very little power control is required.* It is merely necessary to provide adequate power to overcome up-link thermal noise and to assure that the difference between powers in adjacent time slots does not cause serious transient overlap problems.† Normally this is not a serious situation with power variations less than 10 dB, typical transmitter, transponder, and receiver filters, and usual guard times. The down-link power levels can be monitored at a control station. However, this provides very little information about the up-link power levels. The best measure of these levels that can be obtained on the down-link is indirect. If the time slot following a particular transmission is vacant, one can infer the level of the up-link power (very crudely) by observing the transient in the vacant time slot. At best, this is a poor procedure. An improved alternative is to design a satellite telemetry system that transmits the up-link power levels in each time slot down to the central control station. This is rather complex, since the time-slot structure will change with time, thus requiring an adaptive time-gated power meter. Fortunately, for most circumstances it is not necessary to measure (at a central station) the up-link power levels. Power control, if required, can normally be accomplished on a local level by each user adjusting his power until no significant change in error rate occurs. Figure 18 is a diagram illustrating the possible power-control system.

FDMA does require power control to achieve optimum performance. The power-control sensing can be accomplished at a central control station by observing the down-link transmission. Figure 19

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*The power-control problem is complicated, if, as is frequently the case, low-level, in-band, spread-spectrum ranging signals are present. This situation is discussed more fully in another subsection.

†This transient overlap problem is analyzed in a separate subsection.

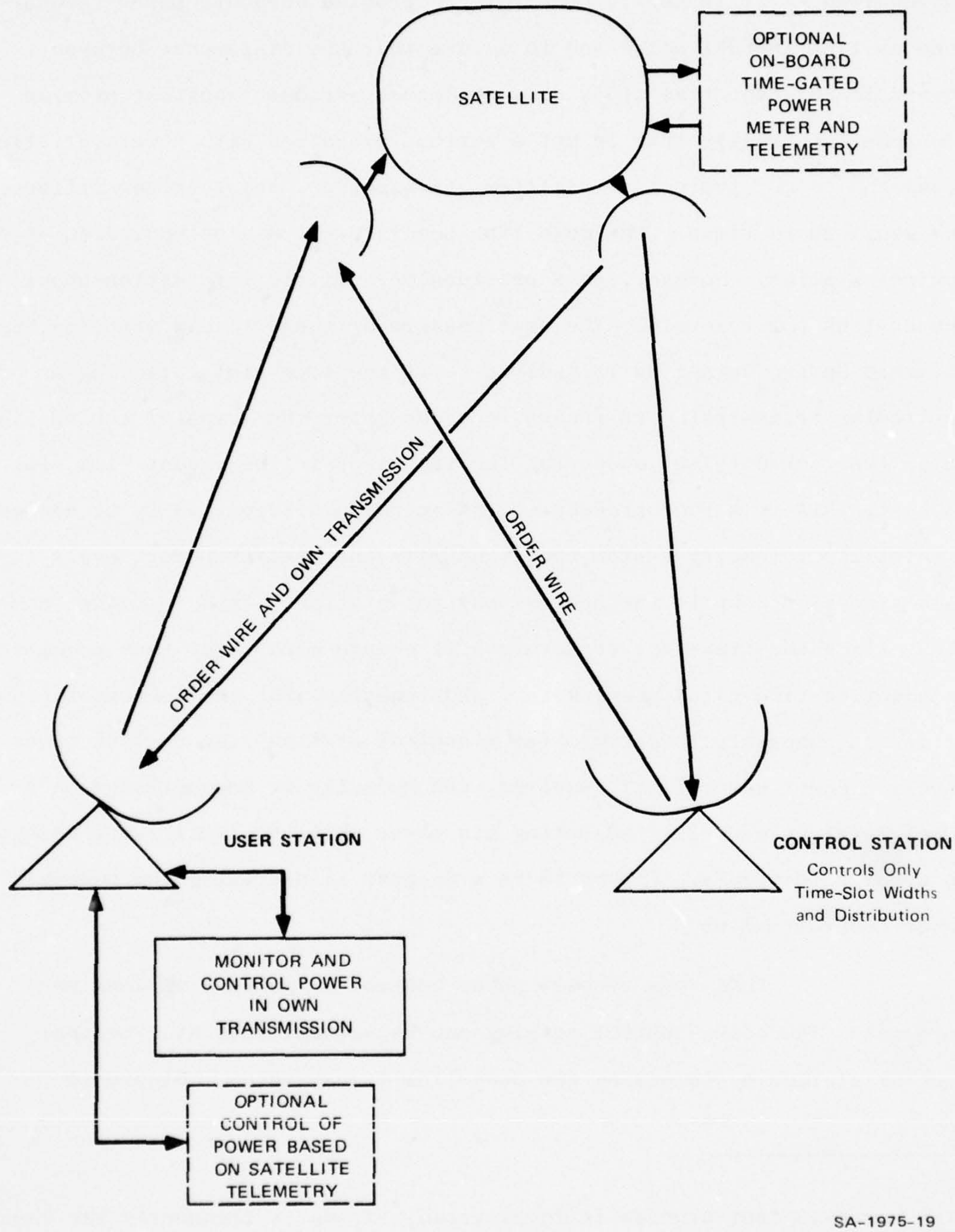

FIGURE 18 DIAGRAM OF POSSIBLE TDMA CONTROL SYSTEM

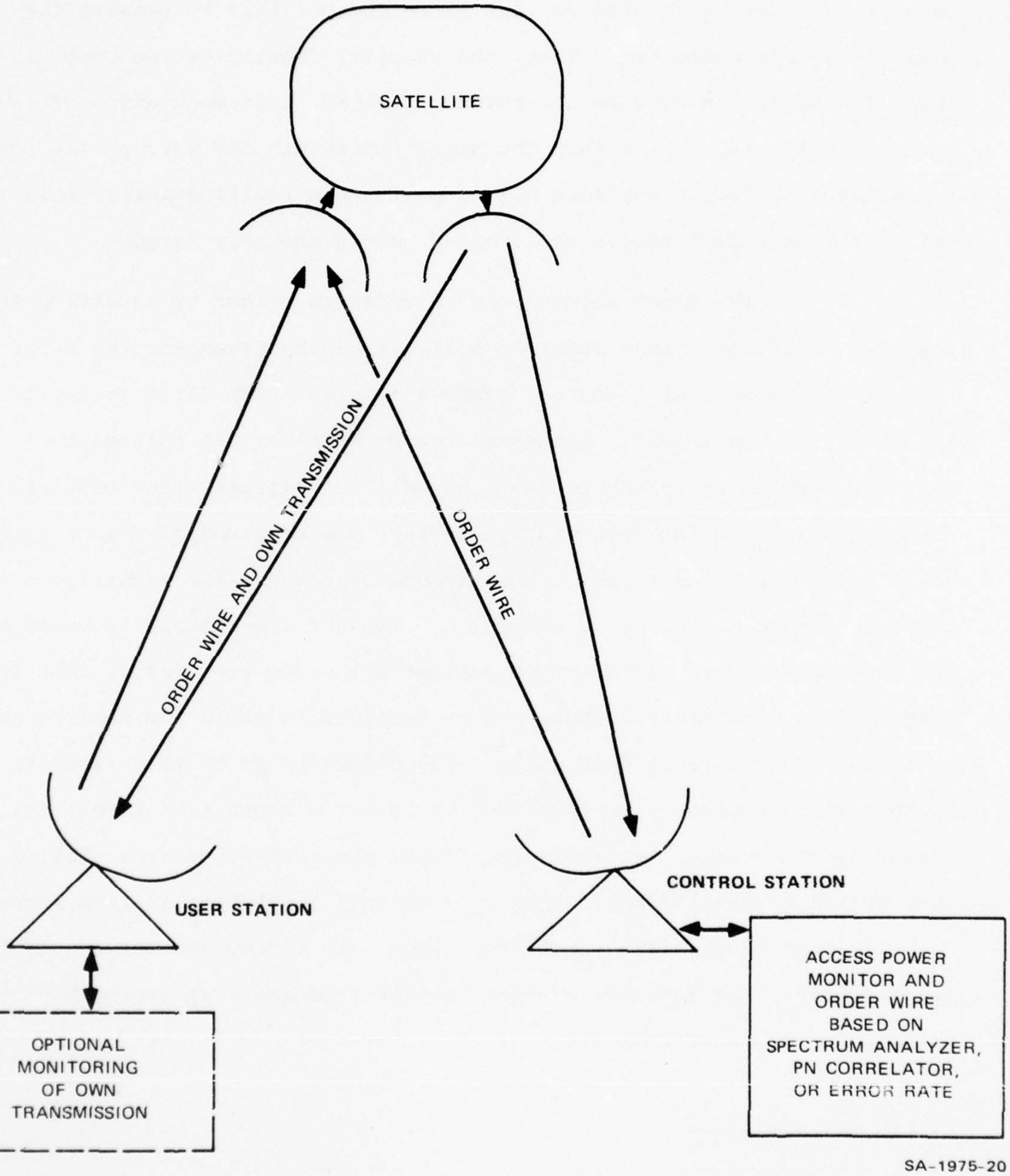

FIGURE 19 DIAGRAM OF POSSIBLE FDMA AND SSMA POWER-CONTROL SYSTEMS

illustrates a possible power-control system for either FDMA or SSMA. The sensing of power levels is complicated by several factors. First, the accesses are data-modulated so that it is not possible to measure the power in a line component. Thus, the spectral density or the power in a specified bandwidth must be determined. Second, intermodulation cross-products will also fall within the signal bandwidth and corrupt the power measurement by indicating more signal power than really exists. This is not a serious effect unless the cross-products are very large.

The power sensing can be effected either by observing the spectral densities with a spectrum analyzer, or by measuring the error rate of each access at a central control station. The first system is the easiest to implement. Advantage can be taken of the existence of digital spectrum analyzers that are directly compatible with computer-input interfaces. The disadvantage is that the measurement also records noise and cross-product power. The second system has the advantage of yielding a much more precise control. Note that the control is based on the true performance criterion of average bit error rate rather than on power. Thus, the control procedure is designed to yield the desired and optimized error rate at each user.* The disadvantage of this approach is that it is necessary to periodically insert a known test signal for measuring error rate. Furthermore, either the central control station must switch error-rate monitoring from channel to channel in time sequence, or there must be multiple receivers. Thus, the second control system is quite complex. It is expected that the spectrum-analyzer approach will be adequate for most FDMA systems.

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*Note that the user G/T's (capacity quotients as well) may differ significantly from those of the central control station. Thus, the desired error rates used in the control algorithm must be transformed to reflect these known differences in G/T.

Power control for SSMA(CDMA) is very similar, with two important differences. First, it is necessary for the power-control algorithm to reflect the effect of access noise (see Section III-A-2-b). Second, power sensing is much more difficult, since a spectrum analyzer cannot resolve the overlapping spectra. Thus, separate receivers for each access (code) are required, or one receiver must be time-shared between the different accesses. The latter approach is complicated by the code-acquisition problem associated with SSMA.* The former approach requires a large amount of equipment. However, it may be necessary for other purposes and consequently be no penalty for a SSMA system.

5. Data-Rate Limitations

TDMA is a very flexible system in many regards. For example, if it is necessary to change data rates it is not necessary to change carrier frequency, retune receivers, and change filter bandwidths. The required changes are simply effected in an all-digital fashion by changing the time-slot size. Thus, on first observation it appears that TDMA is a very flexible system. The purpose of this subsection is to explore some of the limitations on user data rates in a TDMA system.

The burst rates with initial TDMA systems were restricted to be of the form 75.2^n baud. Unfortunately, this meant that the steps (octaves) were rather large and corresponded to a 3-dB change in E/N_o.† A smaller adjustment was desired, so two sets of burst rates were developed. The first set remained of the form $75 \cdot 2^n$ baud, while the second is of the

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*This problem is difficult for initial acquisition only. Thereafter, it should be possible to remember the code time base for each access and thereby greatly speed signal acquisition.

†With this step size it is possible to lose nearly 3 dB in throughput under worst-case circumstances and 1.75 dB on the average [see Figure 20(a)].

form $75 \cdot 3 \cdot 2^{n-1}$. With the intermediate burst rates specified by the 3/2 ratio, it is possible to approximately halve the step in E/N_o.* Thus, for TDMA systems, the burst rates must be of the above form. Typical ranges are from 2.4 to 160 Mbps, depending on the size of receiving ground terminal.

Burst rates are of importance since they determine the transmitted symbol duration and thus have impact on increments in time-slot sizes. Any reasonable TDMA system will contain an integer number of symbols within a time slot. The number of RF bits within one frame must equal the number of baseband bits input at all users (principle of conservation of bits in an information-preserving system). Thus, restrictions on the burst rates imply restrictions on the baseband bit rates. These restrictions may prevent operation at the desired rate or may force operation in an inefficient mode--e.g., dummy bits added to obtain the required rate.

Another limitation on baseband rates is implied by the frame rate. With a frame rate of 1200 fps it is not possible to send data at a rate lower than 1200 baud (2400 baud if quadriphase modulation is used). This restriction can only be overcome if a complex superframe structure is employed that permits low-rate users to skip most frames. For example, a 75-baud user would transmit in every sixteenth frame.† A more complex

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*With this set of step sizes the worst-case loss in throughput is nearly 1.75 dB, and 0.7 dB on the average. Thus, a substantial improvement results [see Figure 20(b)].

†The feasibility of doing this depends on the type of TDMA system and the phase and frequency stability of the link including the terminal equipment. For example, this mode of operation may not be possible with an inertial-type TDMA system operating under conditions of high phase noise. A carrier reconstruction loop with sufficiently long time constant to maintain lock from one burst to the next would be unable to track rapid phase jitter.

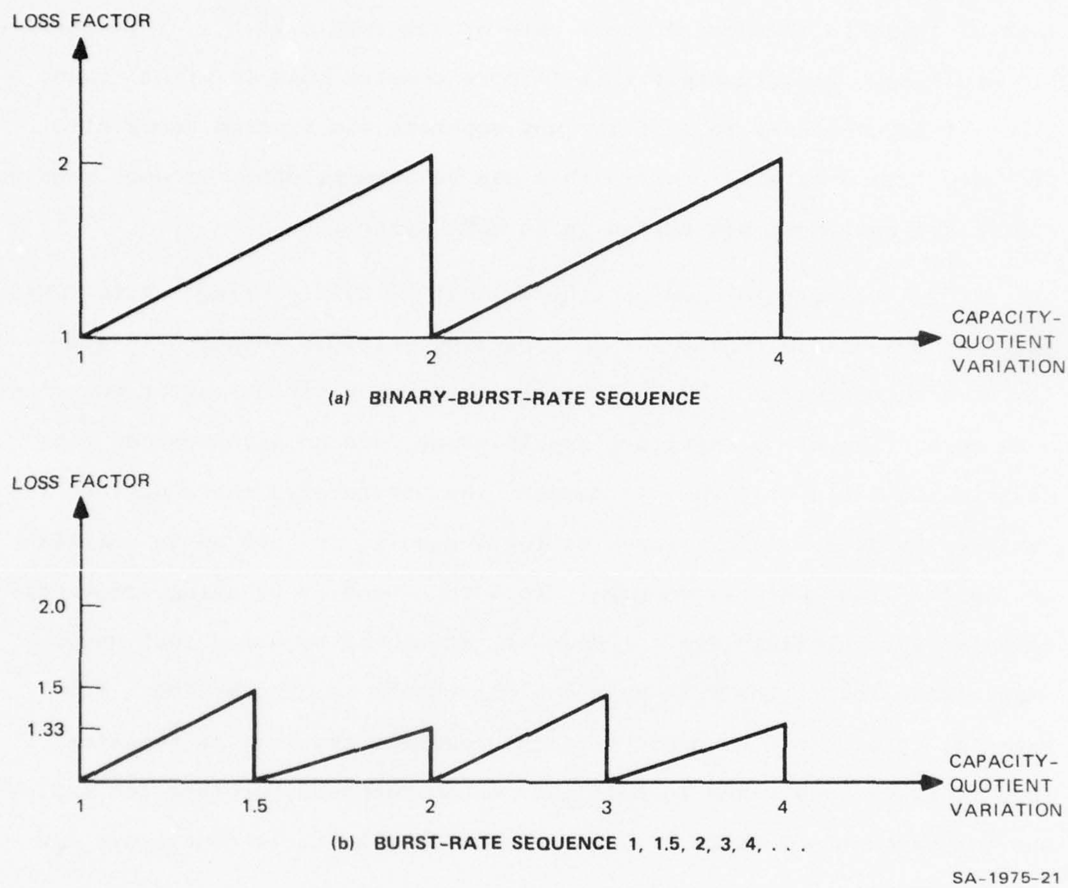

FIGURE 20 BURST-RATE/CAPACITY-QUOTIENT MISMATCH-LOSS FACTOR

receiver is required to accomplish this demultiplexing operation, but it can be accomplished. By contrast, an FDMA simply uses a lower transmission rate and no additional demultiplexing is required.

The fact that not all users have data rates of the form $C \cdot 75 \cdot 2^n$ (where C and n are integers) causes significant problems. If there is to be an integer number of bits in each time slot, then the input streams must be of the above form. In practice, they frequently are not. For

example, 50 kbps secure voice and multiples of 4 kbps (commercial standard) are not of the above form. Thus, it is necessary to increase the rate of each of these to the next highest rate of the form $C \cdot 75 \cdot 2^n$. This involves bit stuffing. Unfortunately this is more complex than it might sound, since it is necessary to identify and separate the stuffed dummy bits from the true data bits. While this can be accomplished, it does require additional equipment not needed in an FDMA system.

A closely related problem is that of multiplexing. With TDMA, the only reasonable choice for multiplexing multiple channels is time-division multiplexing (TDM). However, under many circumstances the channels requiring multiplexing are asynchronous, and an asynchronous time-division multiplexer (ATDM) is needed. Unfortunately, these devices (if designed to accommodate a range of input conditions) are quite complex and costly. An FDMA system might avoid this problem by using frequency-division multiplexing (FDM).* However, it should be noted that under many circumstances the FDMA system might choose to use an ATDM. For example, if multiple channels were intended for the same destination, then it makes good sense to multiplex these channels together and employ one quadriphase modulator rather than building multiple modulators and demodulators.

6. Anti-Jam Capability

The first subsection below compares the AJ performance possible with TDMA, with that achievable with SSMA. The second subsection considers the desirability of mode switching to obtain the required

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*COMSAT's SPADE system is an example of how the asynchronous-clock problem can be avoided through FDM. With SPADE each channel modulates a separate carrier. The complexities of ATDM are avoided at the price of multiple modulators and demodulators.

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performance. The effect of jammer location on mode-switching systems is discussed in the third subsection.

a. Anti-Jam Performance Possible with TDMA

TDMA does not have inherent AJ capability. However a TDMA system can be readily constructed that does have AJ processing capability. Perhaps the simplest form of immunity is obtained by pseudo-randomly selecting different time slots in each frame. This procedure, which certainly presents operational problems (e.g., initial acquisition), prevents a jammer from concentrating its energy on a specific link. If this is the goal of a jammer, then the procedure may be worth the cost and effort. However, a more likely goal of the jammer is to deny any or all signals access to the satellite transponder. In this case, pseudo-random time-slot hopping offers no advantage at a significant price in complexity.

A more effective method is to employ spread-spectrum modulation (as opposed to multiple access) within each time slot. In this case, processing gain can be applied independent of the jammer's goal. If the jammer desires to concentrate on a particular link (time slot), then pseudo-random time-slot hopping can be employed in addition to negate this threat. In fact this mode will be necessary to achieve performance equivalent to that of a SSMA system.*

A TDMA system using spread-spectrum modulation encounters some limitations that are not usually significant with SSMA. If the RF bandwidth is restricted, then the code chipping rate is limited. The AJ

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*Some have argued that the AJ immunity of an SSMA system is reduced due to the presence of other accesses. While this is true, the effect is usually negligible because significant jamming threats have far more power than the other accesses.

immunity is directly proportional to the processing gain, which is the ratio of the code rate to data rate. For TDMA, the burst data rate is the parameter that affects the phase-code processing gain. Thus, for the case of N identical accesses the phase-code processing gain is decreased by the factor N. If the jammer is a CW sine wave, then the amplitude (on/off) code fully recovers this loss. However, since the normal TDMA amplitude code (without time-slot hopping) is so simple, an intelligent jammer could easily match its transmission to a single time-slot burst. In this case the processing gain loss factor of N cannot be recovered. For either case, due to the high burst data rates, the chip-to-bit ratio may fall sufficiently low that poor code tracking will result and further degrade performance.

Another potential disadvantage with TDMA is related to the tradeoff between peak power and pulse length while maintaining a fixed average power. If it were possible to make this tradeoff perfectly, as may well be the case with a properly designed TDMA system, then no disadvantage results. However, if the TDMA system uses peak-power-limited ground transmitters--e.g., transmitters that were designed for continuous operation (with FDMA)--then a significant disadvantage will result. This will occur also when pulse lengths are sufficiently short that it is not possible to increase peak power further. In the former case, the jammer merely needs to compete with one up-link signal at a time. By contrast, with SSMA the jammer must compete with N simultaneous signals. Thus in this case TDMA is at an N-fold disadvantage. Appendix E develops the mathematical basis for the above claim.

b. Mode Switching to Obtain Anti-Jam Capability

The limited RF spectrum tends to dictate against the use of spread-spectrum modulation during normal circumstances. Thus it is

desired to employ spread spectrum only when needed. This implies the need to have parallel equipment or one equipment that can switch its mode of operation. These arguments apply to the use of both TDMA and FDMA in the clear mode.

First, let us consider how one might design one piece of equipment to operate in either a QPSK/FDMA or a PSK/SSMA mode. (A very similar approach could be developed for QPSK/TDMA or PSK/SSMA modes.) A QPSK modulator can be simply modified to produce SSMA merely by mod-2-adding PN codes to each of the data streams (in-phase and quadrature channels). Thus, the modification of the modulator is very simple.

Figure 21 presents a conceptual method of constructing a mode-switching demodulator. As shown, this diagram represents a spread-spectrum demodulator. However, by holding the PN codes fixed (i.e.,

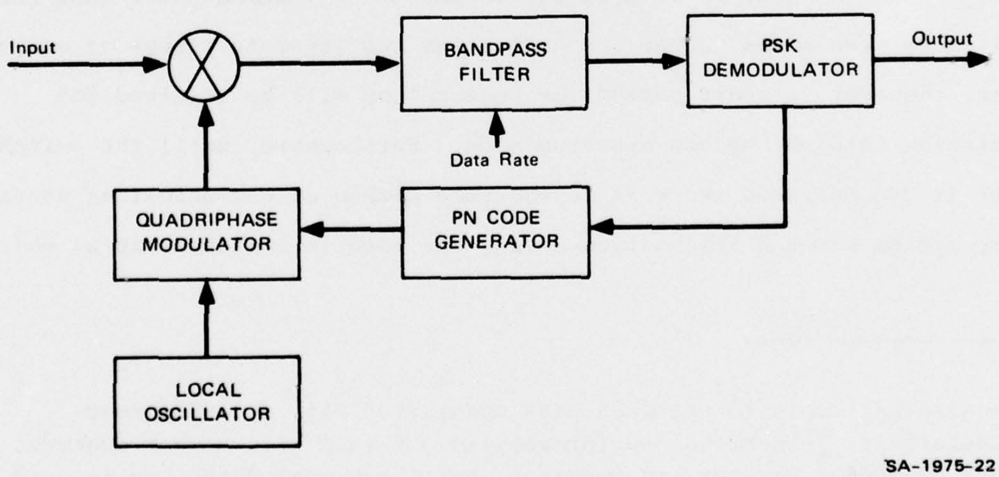

FIGURE 21 BLOCK DIAGRAM OF SPREAD-SPECTRUM DEMODULATOR

clamped to a zero or a 1) the reference signal to the first mixer becomes an unmodulated sine wave. In this case, data demodulation can take place. It should be noted that normally spread spectrum employs binary-data modulation with quadriphase code modulation. Thus, it would be necessary to switch from a binary PSK demodulator in the SSMA mode to a quaternary PSK demodulator in the FDMA mode. This requirement for mode switching will complicate the data demodulator, but it is probably preferable to constructing separate BPSK and QPSK demodulators.*

To permit effective mode switching it is crucial that the PN-code generators be properly timed in advance so that mode switching can occur rapidly. Thus the block diagram of Figure 21 assumes that a portion of the data channel in the clear mode is used to carry timing information to the PN-code generators (if it is not, some other form of synchronization is required). Some acquisition time will still be required, since the timing accuracy of the data channel is typically less than that required by the spread-spectrum equipment.

Thus, it is possible to construct a demodulator that can switch between modes rather than requiring two separate pieces of equipment. However, a short period for acquisition will be required for switching into the spread-spectrum mode. Furthermore, until the switch-over is accomplished there is no reliable method of communicating necessary system command information--e.g., the need to switch to an AJ mode.

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*An alternative is to use QPSK data modulation with the SSMA mode. Theoretically, identical performance is obtained with either coherent BPSK or QPSK. However, in practice, quasi-coherent detection is used since it is necessary to reconstruct a carrier phase reference. QPSK modulation will have greater difficulty performing this function than BPSK. Nevertheless, for a system operating at a reasonable error rate and at a bit rate considerably greater than the carrier-tracking-loop bandwidth, one might expect similar performance in a jamming environment.

By contrast, if the spread-spectrum modulation is always present, a reliable order wire is continuously available to signal the need to switch modes. This approach does require parallel FDMA (or TDMA) and SSMA equipment. However, this seems to be a small price to pay for providing the required system control. Thus, single-equipment mode switching is not recommended as a method of obtaining AJ capability.

c. Effect of Jamming Location on Mode-Switching Systems

Jamming can occur either on the up- or down-links. In general, up-link jamming will affect all the system users,* while down-link jamming will affect only one user. Consequently, with down-link jamming it is undesirable if the jammed access reverts to a full-power and full-bandwidth SSMA format. If it does revert, the SSMA access will seriously interfere with the other accesses. Consequently, three-mode systems have been proposed.

In a three-mode system, down-link jamming (or selective up-link jamming) causes the affected access to use SSMA within its time slot or frequency channel (it is assumed that the data rate is greatly reduced). Thus, in this mode a jammed access does not interfere with other accesses. With non-selective up-link jamming each access reverts to full spread spectrum and mutual interference results. However, under conditions of severe jamming this interference is small compared to the effect of the jammer.

The disadvantage associated with this logical three-mode (clear, down-link jamming, and up-link jamming) system is the complexity associated with the number of modes. Selecting the correct mode can

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*An exception is when the jammer chooses to selectively concentrate its energy on one of the TDMA or FDMA accesses.

represent a significant problem when no reliable order wire is available. It would appear preferable to permit each access to have an additional low-level SSMA signal for order-wire purposes. Thus, in the event of jamming of either type, the affected access would simply remove its clear-mode transmission and use the SSMA signal for a reduced transmission capability. The power level of the signal would only be increased when it was established that up-link jamming was occurring or that it was necessary to permit communication for high-priority users. In effect this resulting system also has three modes. However, the distinction is that this latter system has a reliable order wire present at all times.

C. Frequency-Division Multiple Access

Frequency-division multiple access (FDMA) is not discussed here in the same detail as the other multiple-access techniques, since it is the subject of the major effort of this study contract and has been treated in the reports on the other major tasks of this effort. This subsection is devoted to those areas previously neglected and/or beyond the scope of the original precise analysis effort.

1. Types of FDMA Systems

The types of FDMA systems are restricted to being either QPSK/FDMA or FDMA/SS (i.e., horizontal stacking of spread-spectrum signals) for the jamming environment. It does not appear to be possible to find other major FDMA categories.

2. Multiple-Access Capability

The first subsection below analyzes the effect of adjacent-channel interference on FDMA access noise when heavy transmit and receive filtering is employed. It is found that the access noise can be made negligible by proper choice of filtering. The price to achieve this immunity to adjacent-channel interference (the dominant type of access noise) is a degradation of approximately 1 to 1-1/2 dB due to intersymbol interference. This approximate analysis was performed because it was beyond the scope of the present effort to include the effects of the filtering in the precise computer-evaluation task.* Nevertheless it is believed that filtering effects are significant and must be included if

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*Precise evaluation of intersymbol interference effects is of the same order of difficulty as evaluation of the effect of the transponder nonlinearity on probability of error.

only approximately. Failure to permit the use of transmit and receive filters with an FDMA system will unfairly penalize FDMA due to adjacent-channel interference. This penalty will be particularly severe when there is a large variation in power or power spectral densities.

The second subsection compares the multiple-access performance obtainable by FDMA/SS, with that obtainable by SSMA for the case of identical accesses. It is found that FDMA/SS offers an approximately 9-dB reduction in access noise. However, this effect will not be significant if there is adequate processing gain. The third subsection considers a composite FDMA/SSMA system and compares its performance with that of the FDMA/SS approach--i.e., horizontal stacking of spread-spectrum signals.

a. Adjacent-Channel Interference with FDMA Using Restricted-Bandwidth Transmission

The existence of necessary power or power-spectral-density imbalances can cause significant impact on multiple-access systems. For example, it has been established that for SSMA and significant power imbalance it becomes necessary to assign more power (than predicted on the basis of orthogonal-access power sharing) to weak signals to combat the access noise from the strong signals. This occurs due to the quasi-orthogonal character of all practical SSMA signals.

No such significant effect occurs with TDMA, assuming adequate guard times. However, with FDMA a similar effect occurs due to adjacent-channel interference associated with realizable bandpass filters and their skirt selectivity. It is expected that this nonorthogonal-access effect will be less with FDMA than with SSMA. Several programs were developed to assess the significance of this phenomenon.

The basic situation posed consisted of an adjacent-channel PSK signal interfering with the reception of the desired signal. An expression for the interference power was developed as a function of (1) the carrier offset between the adjacent channels, (2) the signaling rate in the adjacent channel, (3) the power of the interference, and (4) the system filter characteristics. The object was to determine if the interference power could be made sufficiently small that negligible degradation occurred while maintaining high bandwidth utilization for the case of significant imbalances in power spectral density.* The receiving-filter characteristic consists of a cascade of three stages corresponding to RF, IF, and bit-detection filters. The bit-detection filters were assumed to be of the filter-and-sample type, rather than the integrate-and-dump type, since the former provide better performance than the latter for the case of heavily filtered signals.

Program INT1 (see Figure A-6) evaluates the interference power for the case of no transmitter filter, and with receiver-filter pole locations as illustrated in Figure 22. Note that eight poles (referenced to baseband) are specified by coefficients (A1,B1,C1), (A2, B2,C2), and (B3 and C3). For convenience, the program also plots the magnitude squared of the transfer function. Thus, the appropriate choice of the previously mentioned coefficients can be ascertained by comparison of the frequency response function with the desired characteristic (perhaps measured experimentally).

Program INT2 (see Figure A-7) was developed since it became obvious that it was very desirable to suppress sidelobe splatter at the transmitter by appropriate filtering. This program assumes three-pole

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* An additional criterion is that the filtering employed to reduce the adjacent-channel interference, not create significant intersymbol interference.

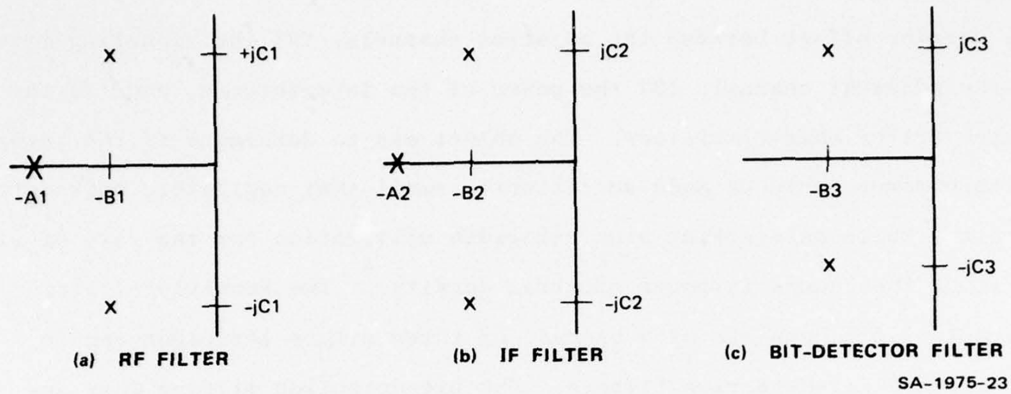

FIGURE 22 POLE LOCATIONS FOR PROGRAM INT1

filtering, specified by coefficients A4, B4, and C4, at the transmitter. In addition it assumes that the bit-detector filter has three poles (see Figure 23) as compared with the two-pole filter of INT1.

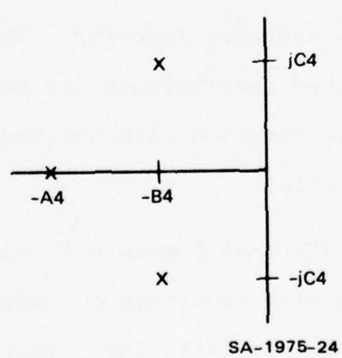

FIGURE 23 POLE LOCATIONS FOR THE BIT-DETECTOR FILTER FOR PROGRAM INT2

Since INT2 does not calculate the receiver frequency response, a separate program PFLT (see Figure A-8) was developed to ascertain that the coefficients Ai, Bi, and Ci for i = 1 to 4 were selected properly to yield the desired transmitter and receiver-frequency-response functions.

Program INT2 is the more useful of the two programs that evaluate the interference power since transmitter filters will be employed with FDMA systems, where bandwidth utilization is very important.

Some examples were evaluated to assess system performance. The normalization selected was such that the symbol duration of the adjacent-channel interference was unity. Thus, the first null in the spectrum was located 2π (angular frequency) from its carrier frequency. Results were obtained for two filter characteristics. The frequency responses for the broad and narrow characteristics are given in Tables 6 and 7, respectively. Note that the broad characteristic places the 3-dB point in the response essentially on the first null of the desired signal. For the narrow-characteristic filter the 3-dB point falls at approximately 80 percent of this value.

Table 6

BROAD-CHARACTERISTIC FILTER

Angular Frequency (rad/s)	Response (dB)
0	0.00
2	-0.24
4	-0.998
6	-2.35
8	-4.46
10	-7.49
12	-11.46
⋮	⋮
20	-32.5

Table 7

NARROW-CHARACTERISTIC FILTER

Angular Frequency (rad/s)	Response (dB)
0	0.00
2	-0.35
4	-1.49
6	-3.66
8	-7.20
10	-12.22
12	-18.4
⋮	⋮
20	-44.9

The interference was assumed to have the same power as the desired signal. For the broad-characteristic filter the channel offset frequency was selected to be 4π rad/s, corresponding to locating the carrier of the adjacent channel at the second null in the spectrum of the desired signal. With this arrangement the main lobes of the two signals are contiguous. The resulting interference level was calculated by Program INT1 to be -9.8 dB. This value is sufficiently high that it has a significant effect on the error rate. Furthermore, if the adjacent-channel signal had greater power (as might be needed to accommodate a low-sensitivity terminal), very poor performance could result.

Similar calculations were performed by INT1 for the narrow-filter characteristic for several channel offset values. These values are presented in Table 8. For the contiguous main spacing the narrower filter reduces the interference power by approximately 4 dB. Even with a spacing as large as 6π rad/s, interference power is quite large and will yield significant degradation if there is a power imbalance. As a result it was concluded that is was necessary to use a filter located at each transmitter to suppress the sidelobe splatter.

Table 8

INTERFERENCE POWER
AS A FUNCTION OF CHANNEL OFFSET

Angular Frequency	Interference Power (dB)
8.5	-6.96
12.5	-14.1
15.0	-16.9
18.8	-19.0

Program INT2 was run for the following case. The transmitter and the receiver RF and IF filters are all identical three-pole Butterworth designs. The 3-dB bandwidth time product is 0.7--i.e., the 3-dB bandwidth occurs at 70 percent of the first-null frequency. With this filter configuration, the power in the interfering signal is suppressed by 25 dB for an offset frequency of 4π rad/s--i.e., contiguous main-lobe channel spacing. This is a very good result for such a tight channel packing (essentially 100-percent bandwidth utilization) and with quite simple filters. Higher-order Butterworth or Chebychev filters can obviously yield better performance with respect to adjacent-channel interference. However, if one chooses to use these simple filters an additional 10 dB in rejection of the adjacent-channel interference can be obtained by increasing the channel spacing from 12.5 to 15.7 (rad/s).

So far no mention has been made of the effect of the filtering on intersymbol interference. Clearly, the adjacent-channel interference can be reduced to any desired degree but at the expense of increased intersymbol interference. Thus, it is necessary to confirm that the intersymbol interference is not excessive for the selected filters.

Precise evaluation of intersymbol interference is a complex problem that requires extensive computation for each filter configuration. Such computation is beyond the scope of the present effort. Rather approximate results based on previous computations will be used to assess the magnitude of the intersymbol interference. The overall transfer function of the filters used in the INT2 example will be approximated by a Chebychev filter of a similar number of poles, and the results of Jones will be used.[9] Program PFLT was run with the identical Butterworth filters (transmitter, receiver RF, and receiver IF) and the overall response was determined. The 3-dB frequency was determined, and this was used with the number of poles (9) to determine the intersymbol interference degradation based on a nine-pole Chebychev filter. This approach yielded an intersymbol interference loss of approximately 1 dB. It may be argued that a cascade of three stages of Butterworth filters should have somewhat poorer performance since it has a poorer skirt selectivity than a Chebychev design. While this is true, it is also true that the use of a three-pole Butterworth bit-detector filter (with BT = 0.7) will yield an improvement over the integrate-and-dump filter assumed by Jones. Consequently it is believed that the degradation due to intersymbol interference is approximately 1 dB and that this value is really quite tolerable.

In conclusion, by appropriate use of practical filtering, the effects of adjacent-channel interference can be made negligible even with power differences of 20 dB. The price paid for the immunity to adjacent-channel interference in the form of intersymbol interference degradation is on the order of 1 to 1-1/2 dB of loss.

b. Comparison of Horizontal and Vertical Stacking of Spread-Spectrum Signals Accessing a Hard Limiter

This subsection makes a quantitative comparison of the performance of horizontal and vertical stacking (i.e., FDMA/SS versus SSMA) of spread-spectrum signals accessing a common hard limiter.

For the purposes of this analysis, it is assumed that there are a large number of accesses of equal power, data rate, and RF bandwidth. Consequently, the power of intermodulation cross products (at the ground) is given by $P_r/8$, where P_r is the total transponder power received at each of the identical ground terminals.

Figure 24 shows how the total available transponder bandwidth is allocated for horizontal stacking.

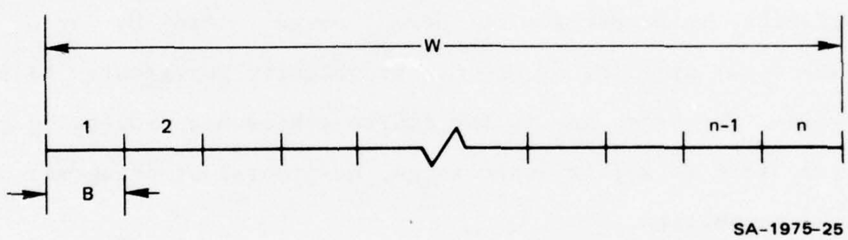

FIGURE 24 HORIZONTAL-STACKING FREQUENCY ALLOCATION

For horizontal stacking, the cross-channel interference power, I_H, is given by*

$$I_H = \frac{b}{B}\,\frac{P_r}{8}\left(\frac{B}{W}\right) = \frac{bn}{W}\,\frac{P_r}{8n} = \frac{P_r b}{8W} \qquad (73)$$

where b is the equivalent-noise bandwidth required to pass the data.

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*It is assumed that the interference power is spread uniformly across the entire bandwidth. This is only approximately true, since the interference power is slightly greater in the center of the band.

For vertical stacking, the cross-channel interference power, I_V, is given by

$$I_V = \frac{b}{W}\left[\frac{7(n-1)\,P_r}{8n} + \frac{P_r}{8}\right] = I_H + \frac{b}{W}\,P_r\left(\frac{n-1}{n}\right)\left(\frac{7}{8}\right) \quad . \tag{74}$$

Thus, $I_V \geq I_H$, and horizontal stacking appears preferable. Consider now the relative degradation of vertical stacking:

$$I_V/I_H = 1 + \frac{7(n-1)}{n} \approx 8 \text{ or } 9 \text{ dB for large n.} \tag{75}$$

Thus, with respect to cross-channel interference, horizontal stacking is about 9 dB superior to vertical stacking. However, if a sufficiently large RF bandwidth is available, the processing gain will be great enough to make the cross-channel interference (in either case) negligible in comparison to thermal noise. Also, by way of comparison, vertical stacking offers the opportunity for greater AJ processing gains. However, due to the finite achievable processing gains in practice, if W is sufficiently large, horizontal stacking may offer an equal AJ capability.

For example, due to power-sharing with the jammer in the satellite transponder, the transmission rate may be limited to the order of 75 bps. The required RF bandwidths with processing gains of 30, 40, and 50 dB are 0.1, 1, and 10 MHz, respectively. Horizontal stacking is possible for a large number of users with the lower values of processing gain and for a modest number with the higher value. Typically, component limitations set an upper bound on the jamming rejection independent of the processing gain. In addition, the processing gain need only make the jamming small in comparison to the thermal noise. Consequently, with practical links, horizontal stacking may offer performance equivalent to that of vertical stacking.

As a result, one may conclude that there is strong merit in considering a QPSK/FDMA system for clear-mode operation. In the advent of jamming, each access would maintain the same carrier frequency but switch to spread-spectrum modulation within its bandwidth. The bandwidths might differ from the clear-mode bandwidths and there might be some spectral overlap in the spread-spectrum mode. Nevertheless, excellent performance could result. One advantage with this FDMA/SS mode of operation is that the system jamming immunity would not be reduced by the direct access noise from the other channels. It would be reduced by the indirect cross-product access noise, which is approximately 9 dB less. This advantage is only significant for modest jamming levels and is negligible for serious jamming threats.

The following subsection considers the advantages of using a composite horizontal-vertical stacking system--i.e., the use of vertical stacking of some signals within a common frequency channel.

c. <u>Composite Horizontal-Vertical Stacking of SSMA Signals</u>

This subsection points out that there exist circumstances in which composite horizontal-vertical stacking may be harmful, as well as situations in which it is helpful.

First, consider the case of an ideal average-power-limited transponder. The obvious motive of nonoverlapping horizontal stacking (frequency offsetting) is to achieve true orthogonality among the n frequency bands, each of which is vertically (coding represented by vertical dimension) stacked with ℓ_i signals, where i = 1, ... n. Thus, a total of $\sum_{i=1}^{n} \ell_i = N$ accesses exist.

As an example, let each of the N accesses have the same power P and let each of the n frequency bands have the same number of

signals $\ell_i = 1 = N/n$. Assume that a transponder bandwidth B is available and that each signal may achieve this bandwidth also.

We now determine the mutual-user interference power at baseband as a function of n. The amount of in-band interference power is $\left(\frac{N}{n} - 1\right) P$. The processing gain is given by B/(nb), where b is the baseband bandwidth. Thus, the interference produces the following power at baseband

$$\left(\frac{N-n}{n}\right) P \cdot \frac{nb}{B} = (N-n)\ P\ \frac{b}{B} \qquad n \leq N \quad . \tag{76}$$

Note that this power is a monotonically decreasing function of n, and the best result is achieved for n = N.

In a jamming environment the processing gain is important, and it may be necessary to restrict $n \ll N$. If this is the case, Eq. (76) is essentially independent of n while the processing gain remains inversely proportional to n. In this case it makes very little sense to increase n. As an example, let there by N = 20 accesses. Going from n = 1 to n = 2 frequency channels reduces mutual-user interference power by the ratio 18/19 or 0.24 dB, while a 3-dB jamming disadvantage is acquired. Thus, horizontal stacking does not appear desirable in a severe jamming environment.

In the case in which some accesses have radically more power than the others, horizontal stacking may be used to advantage. For example, let there be two classes of users that differ in power by 30 dB. Clearly, the processing gain may not be sufficient to give the weak accesses a desirable SNR. However, by allocating the transponder bandwidth to two (one common final amplifier) channels, one for large accesses and one for small accesses, a great improvement can be obtained. While a 3-dB loss in processing gain occurs, this effect is more than

offset by the fact that the strong accesses no longer affect the weak accesses at all, due to orthogonality. Thus, horizontal stacking appears desirable in unbalanced-power circumstances.

Finally, consider the case of a hard-limiting transponder. In this case the horizontal stacking is unable to provide orthogonality because of the transponder nonlinearity. Horizontal stacking will reduce some of the distortion and cross-product terms; however, a reduction in processing gain (that might reduce the effect of these terms) occurs also.

3. Power Control and Reliability

Inadequate power control will cause degradation in FDMA system performance through several mechanisms. First, the power level will be incorrect, since there will be power-control errors. The power level of the i^{th} access at the transponder is proportional to

$$G_i\left(P_1, P_2, \ldots, P_n\right) \cdot P_i \Bigg/ \left(\sum_{j=1}^{n} P_j\right) \tag{77}$$

where P_i is the power in the i^{th} up-link signal, $G_i(\cdot, \ldots, \cdot)$ is the suppression factor associated with the nonlinearity and drive levels, and the remaining factor is the power-sharing effect.

For FDMA systems operating within the normal range, the power control will be accurate to within a few dB.* Under these circumstances the suppression factor is not significantly affected by minor changes in

*Larger variations in power levels will be potentially catastrophic effects and the power-control system must compensate for them. This control system will not work perfectly, and it is the effect of this residual error that we consider here.

levels; consequently, the major variation in power levels will be due to the power-sharing factor. Here there are two effects. The output power in the i^{th} signal is a function of the variations in its up-link power and the other up-link powers.

For the case $P_i \ll \sum_{j=1}^{n} P_j$, one finds that the power in the i^{th} signal is directly proportional to variations in P_i and inversely proportional to variations in $\sum_{j=1}^{n} P_j$. Normally, one expects the denominator to remain very close to its average value for a large number of accesses of roughly equal power. For this case, the variations in the output power will be approximately equal to input-power variations in that signal. If one of the other accesses has a much larger up-link power than the others, the averaging effect will not take place. In this case, the power variation in the desired output signal may be as bad as the sum (in dB) of the variations in the desired up-link signal and the undesired strong signal. Assuming equal power variations for all accesses, this means that the worst-case power-level variation would be doubled. For example, if we controlled the up-link power levels to ± 1 dB, the output-power variation could be as much as ± 2 dB.

For the case

$$P_i \gg \sum_{\substack{j=1 \\ j \neq i}}^{n} P_j$$

the output power in the signal is essentially independent of power variations in its up-link power. However, for this case, variations in P_i inversely affect all other output signals.

The second effect of power-control errors is that the intermodulation cross-products will assume different values. For a well-designed system the cross-products will not have a major impact on system

error rate. Consequently, a small change in cross-product level will have negligible effect. Thus, this effect is not further considered.

Third, power-control errors affect the level of adjacent-channel interference. Again, in a well designed system adjacent-channel interference has a small effect on the system error rate. Consequently, small differences in adjacent-channel interference have negligible effect on system performance. Thus, their effect is not further considered.

Methods of sensing the need for power control for FDMA are discussed in comparison to the problem for TDMA and SSMA in Section III-B-4-b.

As noted in Section III-A-3, the power-control system is not vulnerable to catastrophic failure. That is, unless extremely large power-control errors exist it will always be possible to continue communication but at a reduced transmission rate. While this graceful-degradation feature sounds more desirable than the catastrophic-failure mode of TDMA network timing, such is not necessarily the case. If the threshold of the TDMA network timing system is placed sufficiently far below that of the data system, then the robustness of TDMA network timing will be greater than the robustness of FDMA network-power control.

4. Data-Rate Limitations

FDMA is the most flexible system with respect to data-rate flexibility. Since FDMA does not possess the bandwidth inefficiency of SSMA, the data rate is not so severely limited by the available bandwidth. With FDMA there is no need to precisely interrelate data rates among all accesses. That is, with TDMA, all burst and baseband data rates are coherently related. This arrangement is necessary if the frame time is to be divided into a discrete number of building blocks of fixed duration

(sometimes called quanta) With a TDMA system the guard times will be a discrete number of these quanta. By contrast, no such precise control is required for FDMA.* The guard bands between accesses need not bear any integral relation with the transmission rates. Consequently, FDMA data-rate limitations are set by an individual access rather than the entire network. Thus, there is considerably more flexibility with FDMA.

Limitations implied by a single access are not particularly severe. For example, if the SPADE multicarrier approach is used, virtually any data rate is obtainable provided the data rate on the individual channels is chosen sufficiently low. Furthermore, limitations and complexities caused by asynchronous time-division multiplexers (ATDM) are avoided. The disadvantage with this approach is the large number of modems required.

Use of an ATDM and a single QPSK modulator can significantly alleviate this problem. The only problem becomes one of constructing QPSK modems that will operate at a great variety of data rates. This should be possible. Some attention has been given to developing PSK modems that operate at a great variety of bit rates. Furthermore, some consideration has been given to developing universal modems that will operate at any bit rate. The major equipment difficulty with these approaches is the selection of the proper analog bandpass filters. The digital portion of the modems can readily accommodate variable bit rates.

Thus, one sees that FDMA data-rate limitations are set by equipment complexity and cost rather than network considerations. Consequently, any degree of flexibility can be obtained with FDMA, provided one is willing to pay the cost in equipment complexity.

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*Of course, crude limitations on data rates are present, so that spectral overlap does not occur.

5. Anti-Jam Capability

As with TDMA, FDMA possesses no inherent AJ capability.* However, some capability can be added by hopping frequency slots in a known pseudo-random fashion. This procedure can be effective against a selective jammer but offers no advantage with respect to a jammer whose goal is to deny access to all users of the satellite.

A preferable method of obtaining the jamming immunity is to employ spread spectrum within the frequency slots--i.e., horizontal stacking. The relative advantages and disadvantages of this with respect to SSMA have just been discussed in Sections III-C-2-a and -b, and in Section III-A-5-a, where the effect of access noise on interference immunity was evaluated for SSMA.

The analogy with TDMA is complete with one exception. With TDMA there is the question of the physical capability of trading off duty factor with peak power while maintaining fixed average power. Under some circumstances, this trade cannot be made perfectly and TDMA/SS will be penalized with respect to SSMA. No such problem arises with FDMA/SS. The mode-switching discussion for TDMA applies equally well to FDMA. Thus, a major question exists as to which of the following two systems is best: a QPSK/FDMA system that reverts to an FDMA/SS system when jamming occurs, or a QPSK/FDMA system with low-level SSMA orderwire accesses present at all times. With this second approach, jamming would cause the QPSK/FDMA signals to be eliminated. The advantage of this second approach is that a secure reliable order wire is available at all times to signal the required mode changes. This would appear to be of dominant importance since the advantage of the first approach is simply reduced equipment complexity.

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*This statement refers to the modulation system only and is not intended to apply to the entire communication system. AJ capability can be added to an FDMA or TDMA system by a variety of techniques, such as highly directional antennas.

D. Effect of User Differences on an Ideal Multiple-Access Technique System

In this subsection we develop the impact of a diversity of user types on the performance of an ideal multiple-access technique operating with an idealized power-sharing transponder. The problem is essentially one of resource allocation. The resources are transponder power and bandwidth. Major emphasis is given to the former resource, but the latter is not neglected. The major goal is to show how the satellite throughput (TP) is degraded due to the presence of the user mix. The performance of particular multiple-access techniques can then be compared with respect to this ideal performance.

The harmful effect of a low-capacity terminal on satellite throughput can be best appreciated by considering some simple examples. Here we restrict consideration to the power-limited situation.

For the case of two accesses, each desiring the same baseband data rate, it has been shown that

$$TP = [2/(1 + \Delta Q)]\ TP_{max} \tag{78}$$

where ΔQ is the factor that relates the weaker and strong capacity quotients* and where TP_{max} is the satellite throughput if all the traffic were routed to the stronger terminal. This relationship may be re-expressed in terms of the minimum throughput TP_{min} as

$$TP = [2\Delta Q/(1 + \Delta Q)] \cdot TP_{min} \quad . \tag{79}$$

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*The capacity quotient Q is a measure of the ability of a channel to transmit information. Q = C/kT, where C is the total satellite power referenced to the ground, k is Boltzmann's constant, and T is the system noise temperature.

Note that ΔQ can cause an arbitrarily large degradation in TP with respect to TP_{max}, but can cause only a twofold increase with respect to TP_{min}. Consequently, the introduction of a low-capacity terminal into a satellite system can have an extremely bad effect on satellite throughput. As an example, let $\Delta Q = 9.5$ dB. In this case, $TP = 0.2\ TP_{max} = 1.8\ TP_{min}$. That is, increasing the capacity quotient of one terminal is not nearly as helpful as decreasing the capacity quotient of the other terminal is harmful.

The above results can be readily extended to the case of N accesses each desiring the same baseband data rate. For this case, one finds that

$$TP = \left(N / \sum_{i=1}^{N} \Delta Q_i \right) \cdot TP_{max} \quad . \qquad (80)$$

This result may be restated in terms of the capacity quotients themselves using the following:

$$\Delta Q_i = Q_{max} / Q_i \qquad (81)$$

and

$$TP_{max} = Q_{max} / (E/N_o)^* \qquad (82)$$

where $(E/N_o)^*$ is the required ratio of energy per bit to noise power spectral density to give the desired error rate. The result is

$$TP = \frac{1}{(E/N_o)^*} \cdot \frac{N}{\sum_{i=1}^{N} 1/Q_i} \quad . \qquad (83)$$

Note that second factor of Eq. (83) is analogous to a network of parallel resistors--the resistance values being the capacity quotients. Figure 25 illustrates this circuit. Letting

$$Q_p = 1/\left(\sum_{i=1}^{N} i/Q_i\right)$$

one finds

$$TP = \frac{1}{(E/N_o)^*} \cdot (NQ_p) \qquad . \tag{84}$$

The analogy sheds great insight, since we know from practical experience that one low value of resistance will dominate and control the value of the parallel resistance. The same situation applies for capacity quotients.

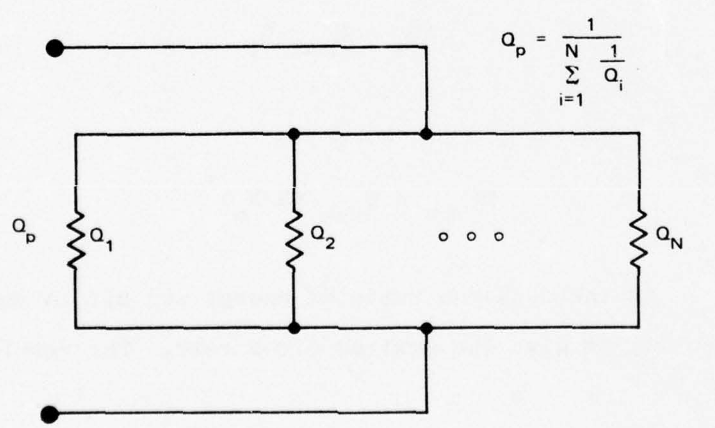

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FIGURE 25 PARALLEL-RESISTOR ANALOGY FOR CAPACITY QUOTIENTS

One should note that due to the factor N in Eq. (84) the analogy is not perfect as compared with respect to parallel resistors. Addition of a large resistance will not increase the parallel resistance. However, because of the factor N an added high-capacity-quotient terminal can increase the satellite throughput.

Consider now the case of unequal capacity quotients and data rates for an arbitrary number of terminals. Here the throughput is given by

$$TP = \left[\sum_{i=1}^{N} (Q_i/E/N_o)_i^*\right] \cdot \left[\frac{(r_i/Q_i)}{\sum_{j=1}^{N} (r_i/Q_i)}\right] \tag{85}$$

where $(E/N_o)_i^*$ is the desired ratio of energy per bit to noise power spectral density in the i^{th} channel, and r_i is the desired baseband data rate in the i^{th} channel. Assuming that all channels desire the same error rate, Eq. (85) may be rewritten as

$$TP/TP_{max} = \sum_{i=1}^{N} (1/\Delta Q_i) \cdot \frac{(r_i/Q_i)}{\sum_{j=1}^{N} (r_i/Q_i)} \quad . \tag{86}$$

Note that the parameter $F_i = r_i/Q_i$ is very fundamental. It is related to the required power in the i^{th} signal. That is, the larger r_i, the more power is needed, and the larger Q_i, the less power is needed.

Since r_i is the desired data rate it is reasonable to consider the case in which the desired rate is tailored to match the intended receiver capacity quotient. That is, $F_i = k$ for all i. For this case,

$$TP/TP_{max} = (1/N) \sum_{i=1}^{N} (1/\Delta Q_i) \quad . \tag{87}$$

Note that this is a different result than for the case of equal data rates for each access [see Eq. (80)]. Consider the two access examples with a 10-dB difference in capacity quotients. For the case of equal F_i, one finds $TP/TP_{max} = 0.55$, while for the case of equal data rates, $TP/TP_{max} = 0.18$. Thus, a substantial advantage, with respect to throughput, can be obtained by tailoring data-rate desires to match the intended capacity quotient.

Let us now determine the tailoring factor $k = r_i/Q_i = F_i$ for all i. Using this philosophy of operation,

$$r_i = Q_i/N \cdot (E/N_o)^* \tag{88}$$

so

$$k = r_i/Q_i = \frac{1}{N} \cdot \frac{1}{(E/N_o)^*} \quad . \tag{89}$$

That is, the selected baseband data rate should be the capacity quotient reduced by the number of accesses and the desired ratio of energy per bit to noise power spectral density.

It should be noted that this mode of operation, while reasonable, does not result in a maximized throughput. The maximized-throughput operation is rather uninteresting. One merely sends all of the baseband data to the receiving terminal with the highest capacity quotient.

The above analyses were based on the assumption of power-limited operation. Under this assumption there is no great difference between the multiple-access techniques. TDMA offers the highest throughput, followed closely by SSMA and FDMA. The very poor bandwidth efficiency of SSMA eliminates it as a serious candidate for high-throughput situations. Consequently, the basic comparison is between TDMA and FDMA. Typically, FDMA is no worse than a few dB from the performance of TDMA.

The large reductions in throughput--i.e., tens of dB's--are due to the differences in user capacity quotients and data-rate requirements.

So far the discussion has ignored the effect of bandwidth limiting. It has been shown in Section III-B-3-b that FDMA can be significantly superior to TDMA for a bandwidth-limited channel with a mixed user population. Thus, it might appear that FDMA should be the recommended multiple-access technique, since in the future bandwidth limitations will be of growing importance. However, this is not really the case, since future systems will not encounter bandwidth limitations substantially before their power limitation. There are several reasons for this situation.

First, the military environment differs greatly from the commercial environment. As greater satellite ERP becomes available due to advanced technology, there will be a reduction in the size of ground terminals, and the existing capacity quotients will thereby be maintained (approximately). The reason for the situation is very obvious. Increased mobility in a military environment is of greater importance than increased capacity quotient. Thus, there is good reason to believe that in the military environment one will always be close to being power-limited.

Second, even if the above is not true, the following effect will occur with any well designed system. A well designed system will be neither greatly power- nor bandwidth-limited. That is, a well designed system tends to operate near the knee of the curve. Thus, for example, as satellite ERP increases there will be a natural tendency to reduce the power efficiency and to operate the power amplifier in a more linear mode, thereby avoiding saturation problems. Consequently, power-efficiency questions may still remain. In addition, as one enters the region of bandwith-limited operation, one will modify the system design. Rather than employing binary or quaternary PSK one will employ higher-order

phase modulation--e.g., 16-ary PSK.* Any such technique for reducing the bandwidth results in reduced power efficiency. In fact the greater the bandwidth reduction the greater the power-efficiency reduction. As a result, with a bandwidth-limited situation a well designed system employs a less power-efficient modulation format. Consequently, operation is never far from being power-limited.

Thus, FDMA does not offer any significant advantage with respect to TDMA under the most likely operating conditions.† Similarly, SSMA encounters the same basic phenomenon of resource sharing. However, due to its very poor bandwidth efficiency and due to the presence of access noise (interference from the other system users), the throughput of a SSMA system is greatly reduced in the presence of a mix of user capacity quotients and data rates. Nevertheless, to a first-order approximation, an SSMA system (of unlimited bandwidth) obeys the same laws of resource allocation as TDMA and FDMA.

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*Alternatively, one might select multilevel FM or perhaps a hybrid AM-PSK system for use with a linearized transponder.

†It should be noted that, for the case of a bandwidth-limited channel and users with large disparity in capacity quotients, the TDMA modem may be much more complex than the FDMA modem. With FDMA it may be possible to use BPSK or QPSK, while for TDMA the high-capacity quotient users will be forced to a higher-order phase modulation. Furthermore, the lower-capacity-quotient TDMA links may not be able to use the higher-order modulation. As a result, the TDMA system may have mixed modulation formats. Such a situation will result in a significant hardware and operational disadvantage for TDMA as compared to FDMA.

E. Summary

In this subsection we compare the multiple-access techniques with respect to the performance criteria of Section II-C.

1. Throughput

In general, the previous analysis sections show that TDMA offers the greatest throughput, followed closely by SSMA and FDMA. The reader is referred to Section IV-B for a comparison on the basis of throughput for a specific test case involving 10 accesses of differing rates and capacity quotients.

2. Tolerance to Inequities Between Receiver G/T and Data Rates

This question basically refers to the ability to accommodate designed and accidental differences in power and power spectral densities. The previous analyses indicate that the SSMA technique is the weakest approach in this regard since it is affected by the access noise that is present with finite processing gain. This effect is sufficiently significant that with large power differences it is necessary to modify the power-control algorithm to account for the access noise.

FDMA can render direct-access noise negligible by employing high-quality transmitter and receiver filters. Thus, the major effects are those of power sharing (also applies to SSMA) and of intermodulation cross-products (IM) (which are a negligible effect for SSMA). By using an optimized frequency plan it is possible to reduce the effects of IM significantly. Consequently, the major problem FDMA encounters is the power-control problem associated with link imbalances. Thus, FDMA can be quite tolerant to inequities and imbalances with a good power-control system.

By contrast, TDMA is bothered by virtually none of these problems. With guard times in excess of 0.5 μs it is possible to make access noise negligible. Thus, TDMA is the most tolerant to inequities and imbalances. However, as noted in Section III-D, the major reduction in throughput is due to the fundamental losses of ideal resource allocation in the presence of user imbalances. The effects described above are secondary in magnitude.

3. Flexibility

FDMA is the most flexible in its ability to accommodate non-standard rates, because there need be no coherent or integral relationship between the rates of different accesses. SSMA has a similar freedom except that limited RF bandwidths and a requirement for a minimum number of code chips per bit restricts the maximum data rates obtainable. Thus, SSMA is inferior to FDMA. TDMA is the most restrictive, due to the constraints among all system users. For example, burst rates must belong to either the chain $2^n \cdot 75$ or the chain $3 \cdot 2^n \cdot 75$. Since there can be no fractional bits transmitted per frame or super-frame and since there are no lost bits (except for overhead functions such as preamble) the baseband data rates are restricted to very specific values.

Flexibility also refers to the ability to meet changing requirements, such as in demand-assignment systems. As noted previously, TDMA has the greatest capability to accept demand assignment. The reason for this is that TDMA can simply reassign time slots. This is a completely digital operation that avoids the requirement for retuning transmitters and changing filters. With TDMA, the operation can be simply and quickly controlled by a computer.

With FDMA, the situation is more complex. However, by employing frequency synthesizers it is possible to permit simple computer

control of the transmitted frequency. The requirement for changing analog RF filters (necessary to reduce adjacent-channel interference) represents a greater difficulty. At present it appears that a rather substantial filter bank would be required to obtain a desirable variety of transmission rates. Consequently, FDMA appears to be more restricted than TDMA with respect to a dynamic demand-assignment situation.

A possible exception to this statement is the SPADE type of FDMA system (in contrast to the QPSK/FDMA system considered in the major portion of this study contract). Here the flexibility is obtained by the multiple-carrier approach. A definite advantage with respect to demand assignment is obtained by this structure. However, the equipment complexity is greatly increased and it is necessary to operate the satellite transponder in a more linear mode to avoid the harmful effects of intermodulation cross-products.

SSMA is less flexible in a static sense, since its peak rate is limited by its poor bandwidth efficiency. With respect to dynamic adjustments it is not troubled by the requirement for retuning carrier frequencies and changing RF filters, as is FDMA. However, the reassignment of resources is somewhat more difficult than for TDMA due to the presence of access-noise effects.

4. Complexity of Operational Doctrine

Operational doctrine complexity is of great interest when it becomes necessary to switch modes of operation. This might be due either to jamming or to changes in user requirement or capabilities. In the former case, SSMA (or a hybrid system employing SSMA) has a clear advantage over FDMA and TDMA.

TDMA has advantages with respect to changes on user capabilities and requirements. Changes are effected in a simple digital fashion

most likely controlled by computer. By contrast, FDMA may require re-tuning of transmitters and changing filters. This may be no particular problem if the changes are relatively infrequent. Furthermore, it is conceivable to automatically control an FDMA system system based on frequency-synthesizer selection of frequencies by computers. The required filtering could be switched in electronically. Thus, the major advantage of TDMA seems to be that it is less sensitive to up-link power variations. As a result, the operational doctrine for this situation is not so complex.

It should be noted that a change in up-link power balance is usually accompanied by a similar problem for the down-link--e.g., rain attenuation applied to both links. TDMA, FDMA, and SSMA are all equally vulnerable to this effect. It becomes necessary to redistribute the resources so the goals are met. TDMA can probably perform this redistribution somewhat more effectively. TDMA need not be so concerned with the effect of path loss (up-link) on achieving this redistribution, since this is accomplished by time-slot durations rather than power levels.

5. Anti-Jam Capability

SSMA obviously has the greatest capability to provide AJ immunity. TDMA and FDMA can both be adapted to provide AJ capability. The disadvantage with this approach is the necessity to employ mode switching at the occurrence of jamming. This arrangement causes undedesirable complexities in the operational doctrine. It would appear preferable to employ a continuously-present, low-level SSMA signal. This hybrid system offers the advantage of a reliable order wire at all times.

As noted above, FDMA and TDMA are inferior to SSMA with respect to jamming immunity. Theoretically, TDMA and FDMA are equivalent in AJ

capability. However, practical limitations on the tradeoff between peak power and duty factor may penalize TDMA, since it may not be possible to obtain the desired peak powers. Consequently, it is estimated that TDMA is inferior to FDMA with respect to AJ capability.

6. Compatibility with Existing Equipment

FDMA is definitely the most compatible with existing equipment. Most existing equipment has been developed for FDMA application either for satellite or line-of-sight type of communication. The only requirement is for flexible QPSK modems to accommodate a variety of bit rates.

SSMA is also compatible with existing equipment. However, it is less so than FDMA due to increased bandwidth requirement in some cases.

TDMA is the least compatible since it requires burst-mode (variable duration as well) operation, while most existing equipment has been developed for continuous operation. Furthermore, the need for network timing can cause difficult timing interfaces with the terrestrial networks. For example, there is the question of how timing and synchronization is established over the ICF links. In addition, special high-rate gated QPSK modems must be developed that are applicable only to TDMA systems and not to conventional terrestrial links.

7. Equipment Costs

Due to compatibility with existing equipment and applicability to conventional communication formats, FDMA is definitely the least costly system to implement. TDMA and SSMA are definitely more costly. At present, it is difficult to estimate which is the most costly. At present, so few TDMA units have been built that it is difficult to accurately forecast their production cost, which is certainly much lower than the

cost of the present development models. SSMA has an advantage in this area since many more units have been produced. Furthermore, the advent of LSI may further reduce the high cost of SSMA modems. The high cost of these modems is in large part due to the requirement for very high AJ immunity. Thus, it is unfair to attribute all of the costs to the multiple-access function. If this requirement is also placed on FDMA and TDMA, their cost will also increase significantly.

It should be noted that FDMA and TDMA are capable of handling much higher data rates than are practical SSMA systems. Therefore, it is reasonable to expect that these equipments might be more costly. It is estimated that if the equipment-cost measure is dollars per bit per second, then the most desirable approach is FDMA, while TDMA is the next most desirable approach. SSMA is estimated to be the most costly in terms of dollars per bit per second.

The above cost comparison is restricted to the multiple-access modems. When terminal equipment cost is included, the cost of FDMA may substantially increase because of a potential requirement for multiple up- and down-converter chains. At present, it appears that multiple converters are required to accommodate multiple-transmission and reception links at each FDMA terminal. This requirement does not hold for other multiple-access techniques.

Further research is required to accurately determine the validity and magnitude of this FDMA terminal cost increase. The first phase of such a study should investigate the feasibility of a simplified up- and down-converter chain that would permit the sharing of the costly (stable and low-phase jitter) components between the multiple chains. The second phase would determine the feasibility of developing "universal" modems that could be tuned to operate at any selected IF. If such modems were feasible, FDMA would not require multiple converter chains.

8. Robustness with Respect to User Errors and Environment Imperfections

All multiple-access techniques require network control to make effective use of the satellite resource. The robustness of this control system with respect to user errors and environmental imperfections is the major factor determining the robustness of the multiple-access technique. With TDMA, the network control problem is principally one of timing, while with FDMA and SSMA, power control is of prime importance. It has been noted previously that network timing is basically an all-or-nothing type of system, while power control possesses more graceful degradation. Thus, potentially, TDMA is less desirable with respect to robustness than is either FDMA or SSMA.

A well designed TDMA system recognizes this potential problem and counteracts it by the following philosophy. First, a carefully conceived failsafe system is incorporated to prevent transmission, should a user lose proper timing. Second, the threshold of catastrophic failure of network timing is placed 10 dB or more below the threshold of catastrophic failure of the data system. Thus, when the timing system fails, it can no longer affect performance. The reader is referred to Section III-B-2-b for a more detailed description of such a system.

Thus, TDMA can be made somewhat more desirable then either FDMA or SSMA with respect to robustness.* The situation can be likened to the choice between analog and digital modulation. Digital modulation does not degrade as gracefully as analog AM or FM signaling. Nevertheless, the performance is sufficiently good and the threshold of catastrophic failure can be made sufficiently low that the digital approach is the preferable one.

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*This slight preference is based on the judgment that it is easier to precisely establish timing control than power control, because it is somewhat more difficult to accurately (with respect to the desired accuracy) measure power than time.

IV COMPARATIVE PERFORMANCE EVALUATION

In this section FDMA, SSMA, and TDMA systems are compared for a specific user configuration consisting of 10 accesses. The quantitative performance comparison is based on total satellite throughput for the three systems.

A. Quantitative Method of Comparing Multiple-Access Techniques

Total satellite throughput has been selected as the quantitative performance measure for the various multiple-access techniques. First, a user mix is selected. This consists of: (1) the number of accesses, (2) the receive down-link capacity quotients, and (3) the data rate desired for each link. It is assumed that all links desire the same error rate (nominally an E/N_o = 10 dB has been assumed to yield a $2.8 \cdot 10^{-6}$ error rate). The desired rates may either under- or over-utilize the satellite power resources. Thus, link calculations based on an ideal power-sharing transponder are performed to assess the resulting E/N_o with optimized power (or time) control. The desired rates of all links are then scaled up or down by the same factor so that the desired error rate is obtained.

The above calculations can be reasonably lengthy, so a simple BASIC program RESA1 (Resource Allocation 1) has been developed to perform these computations. Figure A-9 is a listing of this program. Inputs are the number of accesses, the desired E/N_o, the capacity quotients, and the desired error rates. The program evaluates: (1) the average bit rate obtainable for each access, (2) the percentage of the satellite power devoted to that access, (3) the peak or burst rate (for a TDMA system),

and (4) the total satellite throughput. All of the above quantities are available as factors of the expression for throughput:

$$T = \sum_{i=1}^{n} (Q_i/X) \cdot (R_i/Q_i) / \left(\sum_{j=1}^{n} R_i/Q_i \right) \qquad (90)$$

where Q_i and R_i are the capacity quotient and average bit rate of the i^{th} access, and X is the desired E/N_o.

The output rates from Program RESA1 are then scaled down slightly on the basis of previous experience and these rates together with the other parameters are used as inputs to the SRI FDMA optimization program. The average and individual error rates are then compared with the desired goal. A common scaling factor is then applied to all rates and the optimization program is rerun. This process is repeated until the desired error rate is obtained. The total satellite throughput is then recorded for the FDMA case.

A similar process is pursued for TDMA and SSMA except that the computations are considerably simpler since the statistics of the detection process can be regarded as gaussian and since it is not necessary to evaluate intermodulation cross-products. It should be noted that computer evaluation of the signal-suppression effect is required for the SSMA case since simultaneous accesses are present in a nonlinearity. Fortunately, this information is available from the FDMA computer program. Consequently, no additional computations are required. The resulting error rates for the TDMA and SSMA cases are then determined and all data rates are scaled to provide the desired error rate. The throughputs for the TDMA and SSMA systems are then compared with that for the FDMA system. Thus, a quantitative ranking of the three systems is obtained.

B. Test Case

A user environment was postulated for comparative evaluation of throughput among FDMA, SSMA, and TDMA. A 10-access environment was postulated with the capacity quotients and desired data rates listed in Table 9.* Program RESA determined that the permissible data rates listed in Table 9 should provide a $2.8 \cdot 10^{-6}$ error rate for all accesses under theoretically ideal circumstances. Note that the fifth column is a

Table 9

USER CONFIGURATION

Link Number	Capacity Quotient, Q_i [dB (Hz)]	Desired Rate (bps)	Permissible Rate, r_i (bps)	r_i/Q_i (dB)
1	90	10^5	$6.48 \cdot 10^4$	-42
2	85	10^6	$6.48 \cdot 10^5$	-27
3	80	10^6	$6.48 \cdot 10^5$	-22
4	80	$3 \cdot 10^6$	$1.94 \cdot 10^6$	-17
5	80	$5 \cdot 10^6$	$3.24 \cdot 10^6$	-15
6	90	$2 \cdot 10^7$	$1.297 \cdot 10^7$	-19
7	70	$1.5 \cdot 10^5$	$9.73 \cdot 10^4$	-20
8	85	$5 \cdot 10^6$	$3.24 \cdot 10^6$	-20
9	80	10^6	$6.48 \cdot 10^5$	-22
10	80	10^4	$6.48 \cdot 10^3$	-42

---

*This user configuration was selected not because it represents an optimized demand but because it represents a somewhat unreasonable situation that might occur in a confused operating environment.

measure of the relative signal power that must be given to this link. The two low values (Links 1 and 10) and the high value (Link 5) are underlined.

1. FDMA

The permissible values of Table 9 were used with other outputs of RESA1--e.g., the orthogonal power control values--to choose initial conditions for the FDMA optimization and performance-evaluation routines. The optimization program was then run. Table 10 presents the error-rate results before and after optimization.

Table 10

FDMA ERROR RATES--MAXIMUM THROUGHPUT
(23.52 Mbps)

Link Number	Before Optimization	After Optimization
1	$\underline{4.02 \cdot 10^{-4}}$	$6.56 \cdot 10^{-7}$
2	$2.66 \cdot 10^{-4}$	$1.10 \cdot 10^{-5}$
3	$1.96 \cdot 10^{-4}$	$3.00 \cdot 10^{-5}$
4	$1.23 \cdot 10^{-4}$	$7.18 \cdot 10^{-5}$
5	$6.91 \cdot 10^{-5}$	$\underline{1.55 \cdot 10^{-4}}$
6	$3.19 \cdot 10^{-4}$	$5.69 \cdot 10^{-5}$
7	$1.56 \cdot 10^{-4}$	$4.22 \cdot 10^{-5}$
8	$1.95 \cdot 10^{-4}$	$4.43 \cdot 10^{-5}$
9	$1.92 \cdot 10^{-4}$	$3.00 \cdot 10^{-5}$
10	$2.37 \cdot 10^{-4}$	$5.99 \cdot 10^{-7}$
Average	$2.16 \cdot 10^{-4}$	$4.43 \cdot 10^{-5}$

Using a worst-case analysis (see underlined error rates)* one finds that the loss (compared to gaussian theory) is 2.5 dB before, and 1.9 dB after, optimization. The FDMA program results in a 0.6-dB improvement on a worst-case basis.

Using an average-case analysis one finds that the loss (compared to gaussian theory) is 2.1 dB before, and 1.2 dB after, optimization. In terms of average performance the improvement is 0.9 dB.

The optimization was performed assuming a 185-MHz RF bandwidth, and the resulting satellite throughput is 23.52 Mbps. The resulting bandwidth utilization is 0.127, which is a low value but is consistent with that found for TDMA.

In order to achieve the desired error rate of $2.8 \cdot 10^{-6}$ the satellite throughput must be lowered from 23.52 Mbps. Based on previous experience, all data rates were reduced by a factor of 1.62 with the intention of causing the worst-case error rate to be less than $2.8 \cdot 10^{-6}$. The computer optimization program was rerun with the reduced data rates. Table 11 presents the results. Note that the worst-case error rate (underlined in the table) is $2.15 \cdot 10^{-6}$ which is below and quite close to $2.8 \cdot 10^{-6}$. Thus, it is possible to provide a throughput of 14.55 Mbps while meeting the required worst-case error specification. Note that the average error-rate performance ($5.12 \cdot 10^{-7}$) is considerably better than the specification. One can estimate that it would be possible to increase the throughput (by 0.6 dB) to 16.8 Mbps and still provide the required error-rate performance on an average basis.

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*Note that before optimization the highest error rate corresponds to one of the lowest-value r_i/Q_i accesses. This situation arises due to the presence of adjacent-channel interference and intermodulation crossproducts. After optimization the worst-case error occurs in the highest r_i/Q_i link.

Table 11

FDMA ERROR RATES--REDUCED THROUGHPUT
(14.55 Mbps)

Link Number	Before Optimization	After Optimization
1	$6.8 \cdot 10^{-6}$	$7.01 \cdot 10^{-9}$
2	$5.06 \cdot 10^{-6}$	$1.27 \cdot 10^{-7}$
3	$2.88 \cdot 10^{-6}$	$2.98 \cdot 10^{-7}$
4	$1.51 \cdot 10^{-6}$	$7.92 \cdot 10^{-7}$
5	$6.06 \cdot 10^{-7}$	$\underline{2.15 \cdot 10^{-6}}$
6	$8.07 \cdot 10^{-6}$	$5.55 \cdot 10^{-7}$
7	$2.20 \cdot 10^{-6}$	$4.2 \cdot 10^{-7}$
8	$3.16 \cdot 10^{-6}$	$4.64 \cdot 10^{-7}$
9	$2.88 \cdot 10^{-6}$	$3.08 \cdot 10^{-7}$
10	$4.16 \cdot 10^{-6}$	$7.57 \cdot 10^{-9}$
Average	$3.73 \cdot 10^{-6}$	$5.12 \cdot 10^{-7}$

2. SSMA

The same user configuration was considered for SSMA. It was assumed that each access had potentially differing processing gains and RF bandwidths. With this arrangement the signal-to-interference (access-noise) ratio was set at +16 dB while the SNR was assumed to be +10 dB. Consequently, a 1-dB degradation in error-rate performance resulted. This condition is achieved when $W_i = 4\ Q_{di}$, where W_i is the RF bandwidth of the i^{th} access, and Q_{di} is the effective down-link capacity quotient for the i^{th} access.

Table 12 presents the required processing gain and RF bandwidth to achieve this performance with an SSMA system.* Note that very large bandwidths are required in all cases. The maximum value of 3.9 GHz is obviously not realistic at the present frequency range of 8 GHz. Furthermore, such bandwidths imply phase-modulation rates well beyond practical values. Any SSMA system operating over such a large frequency range

Table 12

REQUIRED RF BANDWIDTH

Link Number	Permissible Rate [dB (bps)]	Processing Gain (dB)	RF Bandwidth
1	48.1	45.4	2.2 GHz
2	58.1	32.1	1.05 GHz
3	58.1	28.0	400 MHz
4	62.9	23.8	470 MHz
5	65.1	20.3	346 MHz
6	71	24.9	3.9 GHz
7	49.9	26.6	45 MHz
8	65.1	25.8	1.23 GHz
9	58.1	28.0	400 MHz
10	38.1	45.6	235 MHz

---

*These values are derived on the assumption of an ideal power-sharing transponder. Consequently, the suppression effects are not included. Fortunately, these are relatively small (approximately 1 dB) and do not have a major impact on the required RF bandwidth, which must be increased approximately 1 dB to account for this suppression.

would undoubtedly have to employ the frequency-hopping approach to spread-spectrum modulation. Consequently, it can be stated that SSMA is not a practical approach for such a user configuration in the foreseeable future. Nevertheless, the performance will be evaluated as if it were possible to implement such a system.

The relative power levels specified by Program RESA1 were then used as inputs to the FDMA computer program. No optimization was performed, but the program was used to evaluate the useful power in each signal component at the limiter output. That is, the suppression loss was evaluated. These numbers were then used to compute the true performance of an SSMA system. Table 13 presents the power-sharing and suppression losses.

Table 13

SUPPRESSION LOSSES

Link Number	Power-Sharing Loss (dB)	Suppression Loss (dB)	Total Loss (dB)
1	31.9	0.9	32.8
2	16.9	0.9	17.8
3	11.9	0.8	12.7
4	7.1	0.5	7.6
5	4.9	0.1	5.0
6	8.9	0.7	9.6
7	10.1	0.7	10.8
8	9.9	0.7	10.6
9	11.9	0.8	12.7
10	31.9	0.9	32.8

The total loss values indicate how much the total capacity quotient is reduced by the presence of other signals. The <u>effective</u> capacity quotient at each access is given by the input capacity quotient (specified as part of the user configuration) less the total loss given in Table 13.

The performance of SSMA can be evaluated in the following fashion. The performance will deviate from that predicted by RESA1 due to the sum of the finite bandwidth loss (1 dB) and the suppression. These values are listed in Table 14 with the effective E/N_o and the resulting error rate based on the Gaussian assumption (which will be very accurate due to the central limit theorem).

Table 14

EFFECTIVE E/N_o AND ERROR RATE

Link Number	Loss (dB)	Effective E/N_o (dB)	Error Rate
1	1.9	8.1	$1.7 \cdot 10^{-4}$
2	1.9	8.1	$1.7 \cdot 10^{-4}$
3	1.8	8.2	$1.4 \cdot 10^{-4}$
4	1.5	8.5	$8 \cdot 10^{-5}$
5	1.1	8.9	$4 \cdot 10^{-5}$
6	1.7	8.3	$1.2 \cdot 10^{-4}$
7	1.7	8.3	$1.2 \cdot 10^{-4}$
8	1.7	8.3	$1.2 \cdot 10^{-4}$
9	1.8	8.2	$1.4 \cdot 10^{-4}$
10	1.9	8.2	$1.4 \cdot 10^{-4}$
Average	1.7	8.3	$1.22 \cdot 10^{-4}$

Using the worst-case performance loss (denoted by the underlining in Table 14) of 1.9 dB we find that the satellite throughput must be reduced from 23.52 Mbps to 15.2 Mbps. Using the average performance loss of 1.7 dB we find that the satellite throughput must be reduced to 15.9 Mbps.

These results indicate that the throughput is only slightly greater for SSMA than for FDMA. For worst-case comparison, however, FDMA does not require the excessive and impractical RF bandwidth of SSMA. It is worth noting for this test case that the bandwidth utilization of SSMA is only $6 \cdot 10^{-3}$ as compared to 0.127 for FDMA. Furthermore, practical FDMA systems employ transmitter and receiver filters not included in the present FDMA analysis. The effect of these filters will be to reduce the adjacent-channel interference (ACI), thereby improving the performance of FDMA with respect to SSMA.*

3. <u>TDMA</u>

The same user configuration was considered for TDMA. The TDMA system selected was based on a 417 μs frame time. A 10-μs time slot is reserved for the central-timing signal. The local-timing signal function is performed by ranging on a local, low-level, continuously present spread-spectrum signal. It is assumed that there are 10 such ranging signals present simultaneously. The total power in these signals is assumed to be approximately 16 dB down with respect to the data signals, resulting in a 1 dB loss. A guard time of 0.5 μs for the 11 signals wastes 5.5 μs in addition to the central-timing slot of 10 μs. This resulting time loss must be added to any preamble loss in each slot.

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*Here it is assumed that the improvement due to reduced ACI will more than offset the degradation due to increased intersymbol interference. It is necessary to establish this hypothesis, and this area is suggested for future study.

It is assumed that the preamble consists of BPSK modulation by a (1, 0) pattern, two symbol durations in length.* This preamble is tracked by a correlation detector that simultaneously synchronizes to carrier-phase and bit-clock timing. It is clear that such a synchronizer will work, since it will average over many preambles, and the SNR for a single preamble must be quite good if a desirable error rate is to be obtained with the data system. Table 15 presents the preamble loss in microseconds for the 10 different links. This table was generated on the assumption that an E/N_o of 10 dB is desired.

The total wasted time is 22 μs, resulting in a frame efficiency of 94.8% or a loss of 0.2 dB. Combining this loss with the ranging signal loss, one finds that the total loss of the TDMA system is 1.2 dB. This means that to achieve the desired error rate of $2.8 \cdot 10^{-6}$, the throughput must be reduced from 23.52 Mbps to 17.8 Mbps.

This value is somewhat higher than the worst-case and average values for FDMA of 14.55 and 15.8 Mbps, respectively. However, it must be recognized that the FDMA system can be improved by the use of transmitter and receiver filters. Furthermore, the FDMA system is considerably simpler (e.g., the highest FDMA data rate is no more than 13 Mbps, while the TDMA system requires 100-Mbps operation), and is more compatible with existing equipment.

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*This preamble is required to identify the start of burst for these circumstances in which the network-timing-system accuracy, while adequate for the guard times employed, is not sufficient to define the first symbol in the burst with great enough reliability. Such bit-slippage possibilities are normally covered by a separate (from error rate) specification commonly referred to as bit-count integrity. The preamble serves the start-of-burst-symbol function that permits this specification to be met. In addition, the preamble is most useful in providing a deterministic waveform for the carrier and bit-sync loops to track even in inertial-mode systems.

Table 15

PREAMBLE DURATION

Link	Preamble Duration (μs)
1	0.04
2	0.13
3	0.4
4	0.4
5	0.4
6	0.04
7	4
8	0.13
9	0.4
10	0.4
Total	6.3

The bandwidth assumed for this TDMA system evaluation is 185 MHz. Note that this amount of bandwidth is required if the highest burst-rate transmissions are not to be significantly degraded with respect to error-rate performance. Links 1 and 6 have capacity quotients of 90 dB(Hz), and consequently will be operated at a 100-Mbps burst rate. A 185-MHz passband very nearly encompasses the spectrum to the second null in the envelope (assuming QPSK data modulation). Negligible degradation due to intersymbol interference occurs for this case.

The bandwidth utilization assuming a 23.52-Mbps throughput is 0.127. The bandwidth utilization based on a bit error rate of $2.8 \cdot 10^{-6}$ is reduced to 0.096.

4. Summary

Table 16 summarizes the comparison of throughputs for the three multiple-access techniques.

Table 16

COMPARATIVE THROUGHPUTS

System	Throughput (Mbps)	Case
Ideal	23.52	All channels
FDMA	15.8	Average channel
	14.55	Worst case
SSMA	15.9	Average channel
	15.2	Worst case
TDMA	17.8	All channels

It should be noted that the above comparisons have neglected two important losses. First, imperfections in power control will result in a loss for both FDMA and SSMA. Second, the fact that TDMA burst data rates are restricted to rather large step sizes results in a loss for TDMA. It is difficult to precisely estimate these losses (as we shall see). However, it is reasonable to guess that a power-control loss of 0.5 dB might be typical for FDMA and SSMA, while a 0.7-dB nonmatched-burst-rate loss might occur for TDMA.

The difficulty in precisely estimating the power-control loss for FDMA can be best appreciated by attempting to perform the calculation in an approximate fashion. First, assume an ideal power-sharing

transponder. Next, consider the actual up-link powers, P_i, to be random variables with a uniform distribution about the mean $\overline{P_i}$, which is the desired value. The range is determined by the uncertainty in power control--perhaps ± 1 dB. Estimating this value is a difficult problem itself.

Assume that the power-control errors are independent random variables. In this case, the average proportion of the transponder power that is devoted to the i^{th} access is given by the expected value

$$E\left\{ P_i / \sum_{j=1}^{n} P_i \right\} \tag{91}$$

where the nominal value is given by

$$\overline{P_i} / \sum_{j=1}^{n} \overline{P_j} \tag{92}$$

where the overbar denotes the mean value. The difference between these two numbers is the average power-sharing loss.* Evaluation of true loss is complicated by the complexity of the density function and the number of variables involved--i.e., the n mean values and the parameter n. Here we assume that all accesses have the same power-control accuracy. The mean loss could definitely be calculated, but it is a very complex computation that is beyond the scope of the present effort. Furthermore, the above computation does not give the result truly desired, although it may be a reasonable approximation.

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*Based on averaging with respect to powers (not over error rates).

The problem with the above approach is that it deals with powers and not error rates. Due to the nonlinear relation between power and error rate, the average loss computed on the basis of powers is not the same as the average loss based on error rates. Furthermore, if one adopts a worst-case-channel-comparison philosophy, then it becomes necessary to evaluate the loss with respect to this worst channel. Such a computation is indeed complex. Perhaps the best solution might be by Monte Carlo simulation using the present computer programs and permitting errors in the power levels. This is clearly on extensive effort that cannot be accomplished within this task. Consequently, we make an estimate that the power-control loss will be about 0.5 dB.

Estimation of the loss due to mismatched (with respect to capacity quotient) burst rate is also complex, since each link has a different capacity quotient and a different rate. Consequently, it is not clear which loss to use or how the losses might be averaged (possibly weighted by the data rate?). Consequently, the mismatched-burst-rate loss is estimated to be its average value based on a uniform distribution over a two-to-one rate range. It is assumed that burst rates can be selected according to the following sequence: 2, 3, 4, 6, 8, 12, 16, ... The loss estimate is then 0.7 dB.

Table 17 summarizes the comparative throughputs with these losses included.

Table 17

COMPARATIVE THROUGHPUTS WITH POWER-CONTROL AND BURST-RATE LOSSES

System	Throughput (Mbps)	Case
Ideal	23.52	All channels
FDMA	14.0	Average channel
	12.9	Worst case
SSMA	14.1	Average channel
	13.5	Worst case
TDMA	15.05	Average channel

V CONCLUSIONS

On the basis of the previous analyses one can conclude that TDMA and FDMA offer comparable performance for a mix of user G/T's and data rates. This result is obtained on the assumption that the same satellite bandwidth and power are available for both techniques. This result may appear surprising at first, since TDMA does not encounter the power-sharing and intermodulation cross-products of FDMA. However, the result may be explained heuristically in the following fashion.

First, the mix in user G/T's causes the bandwidth utilization of a TDMA system to be reasonably low.* By itself this fact is neither good nor bad. What it means is that the required RF bandwidth is sufficiently large that an FDMA system with the same throughput can avoid most of the intermodulation cross-products. Thus, the FDMA system is not seriously degraded by this effect. In addition, a TDMA system may be expected to encounter some degradation from the presence of the local-timing signals. Thus, the TDMA approach has a roughly compensating loss.

The power-control loss of an FDMA system might be estimated to be approximately 0.5 to 1.0 dB. There is no significant power-control loss in most TDMA systems. However, there is an equivalent loss with TDMA. In a TDMA system it is not possible to perfectly match burst rates with capacity quotients. A reasonable estimate of the practical loss is 0.7 dB. Due to their greater flexibility, FDMA systems can effectively

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*The high-capacity-quotient user has the same effect on bandwidth utilization that the low-capacity-quotient user has on throughput in the power-limited case.

avoid this loss. Consequently, with respect to throughput, TDMA and FDMA are roughly equivalent (to within a few decibels).

By contrast, SSMA is seriously hampered due to its bandwidth inefficiency. Large differences in user G/T's and data rates force the required RF bandwidths to be excessive. Consequently, SSMA cannot be given serious consideration as a technique for high-data-rate transmission with a mix of user types. However, it must be noted that SSMA is the only multiple-access technique with inherent AJ capability. Even with respect to this important performance criterion, SSMA is not without significant failing. Without the use of a complex processing transponder, SSMA is degraded by the power-sharing effect in the satellite transponder. Thus, SSMA is not nearly as effective as theoretically possible. Nevertheless it may provide the required order-wire capability for "last-ditch" operation. High-gain directional satellite antennas might provide an equally or more effective solution to this problem and should not be neglected as a possible competitor or addition to an SSMA order-wire system.

So far the discussion has concentrated primarily on quantitative performance measures such as throughput. While such measures are important, they are by no means the only significant measures. Perhaps of greater importance is the robustness of network control. Here a fascinating choice develops. With FDMA (and SSMA) the primary problem is network power control, while with TDMA the significant problem is network timing. The difficulty with TDMA is that the control system is vulnerable to catastrophic failure. However, by careful initial system design the occurrence of this can be made very improbable. By contrast, FDMA degrades gracefully, but significant degradations occur much sooner with respect to small errors in network control. Consequently, FDMA is inferior with respect to user small errors, but TDMA is inferior with respect to user large errors. Thus, under most circumstances TDMA is

preferable. Under unusual circumstances FDMA may more effectively maintain minimal communication.

Additional performance criteria are equipment cost, simplicity, and flexibility. With respect to these criteria, FDMA is equivalent or superior to TDMA. Consequently, it would appear that the FDMA approach, since it yields technical performance comparable to that of TDMA, should be given serious consideration as a most promising technique for military satellite communication.

It must be cautioned that the above conclusions are reached on the basis of a very small study effort (Task 4c of five tasks) and a limited number of examples.† Thus, the conclusions must be regarded as preliminary. They merely indicate that FDMA is worthy of more serious consideration as a multiple-access technique for military satellite communication. Detailed evaluation of FDMA capability must be conducted in subsequent

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*As noted in Section III-E-7, depending on assumptions, FDMA equipment may be substantially more costly than TDMA equipment, because of the potential requirement for multiple up- and down-converter chains for the former system.

†For example, if all users have equal-capacity quotients, TDMA will have a high bandwidth utilization. In this case it will be difficult for an FDMA system (using the same RF bandwidth) to avoid the cross-products. In this case, the FDMA system may be expected to be inferior to a TDMA system. At present, it is difficult to accurately estimate the inferiority of FDMA since the present computer program does not include the filters present in a practical system. As a result, the present analysis masks the basic losses due to IM cross-products with the losses due to adjacent-channel interference (ACI). Consequently, this task of the present study does not report on such a comparison between FDMA and TDMA.

work to verify the present conjectures based on preliminary and approximate analyses. For example, it will be necessary to evaluate the effects of adjacent-channel interference and intersymbol interference as a function of different system filters employed in practical FDMA systems. This aspect is of great importance for high-bandwidth-utilization systems anticipated in future operation. In addition, other important factors such as AM/PM conversion losses should be included in the analyses.

Appendix A

COMPUTER PROGRAM LISTINGS

```
>LØAD SSMA
>LIST
010 PRINT"THIS PRØG DET BW NEEDED FØR N SSMA SIGS"
011PRINT"N=DESIRED NUMBER ØF SIGNALS"
012PRINT"B=BIT RATE ØF EACH USER"
013PRINT"X=REQUIRED E/NØ"
014PRINT"Q=DØWNLINK CAPACITY QUØTIENT"
020PRINT "ENTER N, B, X"
030INPUT N,B,X
040PRINT"ENTER Q(J)"
050FØR J=1 TØ 4
060INPUT Q(J)
070NEXT J
075 DIM W(1:10,1:200)
080FØR K=1 TØ N
085IF K=1 THEN 165 ELSE 090
090W(1,K)=K
100FØR J=1 TØ 4
120N(J+1)=(K-1)*B
130D(J+1)=(1/X)-K*B/Q(J)
135IF D(J+1)<1E-04 THEN 136 ELSE 140
136W(J+1,K)=1E20
137GØ TØ 145
140W(J+1,K)=N(J+1)/D(J+1)
145NEXT J
155PRINT W(1,K),W(2,K),W(3,K),W(4,K),W(5,K)
165NEXT K
170PRINT"K","W1","W2","W3","W4"
171PRINT
172PRINT
175FØR J=1 TØ 4
180FØR K=1 TØ N
185Y=Q(J)/(B*K)
190IF Y<X THEN 212 ELSE 210
210NEXT K
211GØ TØ 220
212PRINT J,K-1, ,W(J+1,K-1),(K-1)*B/W(J+1,K-1)
215GØ TØ 230
220PRINT J,K,W(J+1,K),K*B/W(J+1,K)
230NEXT J
240PRINT"J","K","W","BU"
241PRINT"K=MAX NØ ØF PØSSIBLE SIGNALS"
242PRINT"W=REQUIRED RF BANDWIDTH"
243PRINT"BU=BANDWIDTH UTILIZATIØN"
250END
```

FIGURE A-1 LISTING OF PROGRAM SSMA

```
>LØAD SSPC
>LIST
010PRINT"EFFECT ØF ACCESS NØISE ØN PØWER CØNTRØL"
020PRINT"ENTER X,Y,P1,P2,K"
030INPUT X,Y,P1,P2,K
040A1=X*P1
050A2=X*P2
055L=K*Y
060G1=K*(L↑2-1)
070G2=(L*P2+(2*L↑2-1)*P1)
080G3=Y*P1*(P2+L*P1)
090F=SQR(G2↑2-4*G1*G3)
110C2=(-G2-F)/(2*G1)
130C1=Y*(P1+K*C2)
140PRINT A1,A2,A1/A2
150PRINT "A1","A2","A1/A2"
155PRINT"ØRTHØGØNAL PØWER CØNTRØL"
160 A3=A1+A2
162C3=C1+C2
164S=10*LØG10(C3/A3)
180PRINT C1,C2,C1/C2
190PRINT "C1","C2","C1/C2"
195PRINT"QUASI-ØRTHØGØNAL PØWER CØNTRØL"
196PRINT S
197PRINT"ACCESS LØSS IN DB"
200END
```

FIGURE A-2   LISTING OF PROGRAM SSPC

```
>LØAD SSPC1
>LIST
010PRINT"EFFECT ØF ACCESS NØISE ØN PØWER CØNTRØL"
020PRINT"ENTER X,P1,P2,K"
030INPUT X,P1,P2,K
040A1=X*P1
050A2=X*P2
055A3=A1+A2
060N1=P1+K*A3
070D1=P1+P2+2*K*A3
080D=N1/D1
090B1=D*A3
100B2=A3-B1
110Y=B1/(P1+K*B2)
140PRINT A1,A2,A1/A2
150PRINT "A1","A2","A1/A2"
155PRINT"ØRTHØGØNAL PØWER CØNTRØL"
160PRINT B1,B2,B1/B2
170PRINT"B1","B2","B1/B2"
195PRINT"QUASI-ØRTHØGØNAL PØWER CØNTRØL"
200PRINT Y,-10*LØG10(X/Y)
205PRINT "E/NO","ACCESS LØSS IN DB"
210END
```

FIGURE A-3 LISTING OF PROGRAM SSPC1

```
>LØAD SSPC2
>LIST
005PRINT"ACCESS LØSS AND PØWER CØNTRØL"
010PRINT"ENTER W"
012INPUT W
015PRINT"ENTER N1,N2,B1,B2,X,Y"
016INPUT N1,N2,B1,B2,X,Y
017A1=X*B1*N1
018A2=X*B2*N2
20A3=A1+A2
040D1=B1/W
050D2=Y*D1
060G1=D1*((D2)↑2-1)
070D3=D2*B2
080D4=2*N1*D2
090G2=D3*(D4+N2)-N1*B1
100D5=Y*N1*D1*B2
110D6=N1*D2
120G3=D5*(N2+D6)
130F=SQR(G2↑2-4*G1*G3)
140C2=(-G2-F)/(2*G1)
150C1=Y*B1*(N1+C2/W)
160C3=C1+C2
170PRINT A1,A2,A3
175PRINT"A1","A2","SUM"
180PRINT C1,C2,C3
185PRINT"C1","C2","SUM"
190S=10*LØG10(C3/A3)
195PRINT S
200PRINT"SSMA ACCESS LØSS IN DB"
205END
```

FIGURE A-4 LISTING OF PROGRAM SSPC2

```
>LØAD SSW1
>LIST
010PRINT"THIS PRØG EVAL RF BW AS A F'N ØF RATES & Q'S FØR SSMA"
020PRINT "ENTER NUMBER ØF ACCESSES"
030INPUT M
040PRINT"ENTER DATA RATES"
045DIM B(1:M)
050FØR I=1 TØ M
060INPUT B(I)
070NEXT I
080PRINT"ENTER CAPACITY QUØTIENTS"
085DIM Q(1:M)
090FØR I=1 TØ M
100INPUT Q(I)
110NEXT I
120PRINT"ENTER ØUTPUT SNR"
130 INPUT X
160DIM W(1:M)
170C1=0
180FØR J=1 TØ M
190F=B(J)/Q(J)
200C1=C1+F
210 NEXT J
220FØR I=1 TØ M
230N=Q(I)*C1*X*(1-B(I)/(C1*Q(I)))
240D=1-X*C1
250W(I)=N/D
260NEXT I
270W1=W(1)
280FØR I=2 TØ M
290IF W(I)>W1 THEN 300 ELSE 320
300W1=W(I)
310GØ TØ 330
320W1=W1
330NEXT I
340E1=0
350FØR I=1 TØ M
360E1=E1+B(I)
370NEXT I
380B1=E1/W1
390PRINT W1,B1
400PRINT"W MAX","BANDWIDTH UTILIZATIØN"
410FØR I=1 TØ M
420PRINT W(I)
430NEXT I
440END
```

FIGURE A-5 LISTING OF PROGRAM SSW1

```
>LØAD INT1
>LIST
010PRINT "THIS PRØG EVAL INT PØW FRØM PSK SIG"
018PRINT "ENTER A"
019INPUT A
020PRINT"ENTER A1, B1, C1"
025INPUT A1,B1,C1
030PRINT"ENTER A2, B2, C2"
035INPUT A2,B2, C2
040PRINT "ENTER B3, C3"
045INPUT B3,C3
050 PRINT "ENTER WO,T,D,P,N"
055INPUT WO,T,D,P,N
060L1=T/2
061 L2=(P*T)/(2*PI)
070N1=((B1↑2-C1↑2)↑2+4*(B1*C1)↑2)*(A1↑2)
075N2=((B2↑2-C2↑2)↑2+4*(B2*C2)↑2)*(A2↑2)
080N3=(B3↑2-C3↑2)↑2+4*(B3*C3)↑2
081D4=A1↑2
082E1=B1↑2-C1↑2
083F1=4*(B1*C1)↑2
084D5=A2↑2
085E2=B2↑2-C2↑2
086F2=4*(B2*C2)↑2
087E3=B3↑2-C3↑2
088F3=4*(B3*C3)↑2
089R=0
090FØR K=1 TØ N
100W=(K-(N/2))*D
101W2=W↑2
110D1=(W2+D4)*((W2+E1)↑2+F1)
120D2=(W2+D5)*((W2+E2)↑2+F2)
130D3=(W2+E3)↑2+F3
140M=N1*N2*N3/(D1*D2*D3)
150W1=W-WO
151X=W1*L1
152Z=ABS(X)
153IF Z<0.01 THEN 154 ELSE 160
154S=L2*(1-(X↑2)/6)↑2
155GØ TØ 170
160S=L2*(SIN(X)/X)↑2
170R=R+S*M*D
172Y=KMØD10
175IF Y=0 THEN 176 ELSE 188
176IF A=1 THEN 187 ELSE 188
187PRINT K,W,10*LØG10(M),R
188NEXT K
189IF A=1 THEN 190 ELSE 195
190PRINT"K,W,10LØGM,R"
195PRINT "R,R/P,10LØG(R/P)"
200PRINT R, R/P,10*LØG10(R/P)
205END
```

FIGURE A-6 LISTING OF PROGRAM INT1

```
>LØAD INT2
>LIST
010PRINT "THIS PRØG EVAL INT PØW FRØM RSK SIG"
018PRINT "ENTER A"
019INPUT A
020PRINT"ENTER A1, B1, C1"
025INPUT A1,B1,C1
030PRINT"ENTER A2, B2, C2"
035INPUT A2,B2, C2
040PRINT "ENTER A3,B3,C3"
045INPUT A3,B3,C3
046PRINT "ENTER A4,B4,C4"
047INPUT A4,B4,C4
050 PRINT "ENTER W0,T,D,P,N"
055INPUT W0,T,D,P,N
060L1=T/2
061 L2=(P*T)/(2*PI)
070N1=((B1↑2-C1↑2)↑2+4*(B1*C1)↑2)*(A1↑2)
075N2=((B2↑2-C2↑2)↑2+4*(B2*C2)↑2)*(A2↑2)
079N4=((B4↑2-C4↑2)↑2+4*(B4*C4)↑2)*(A4↑2)
080N3=((B3↑2-C3↑2)↑2+4*(B3*C3)↑2)*(A3↑2)
081G1=A1↑2
082E1=B1↑2-C1↑2
083F1=4*(B1*C1)↑2
084G2=A2↑2
085E2=B2↑2-C2↑2
086F2=4*(B2*C2)↑2
087E3=B3↑2-C3↑2
088F3=4*(B3*C3)↑2
089G3=A3↑2
090G4=A4↑2
091E4=B4↑2-C4↑2
092F4=4*(B4*C4)↑2
098R=0
099FØR K=1 TØ N
100W=(K-(N/2))*D
101W2=W↑2
110D1=(W2+G1)*((W2+E1)↑2+F1)
120D2=(W2+G2)*((W2+E2)↑2+F2)
130D3=(W2+G3)*((W2+E3)↑2+F3)
140W1=W-W0
142W3=W1↑2
143D4=(W3+G4)*((W3+E4)↑2+F4)
144M2=N4/D4
145M1=N1*N2*N3/(D1*D2*D3)
151X=W1*L1
152Z=ABS(X)
153IF Z<0.01 THEN 154 ELSE 160
154S=L2*(1-(X↑2)/6)↑2
155GØ TØ 170
160S=L2*(SIN(X)/X)↑2
170R=R+S*M1*M2*D
172Y=KMØD10
175IF Y=0 THEN 176 ELSE 188
176IF A=1 THEN 187 ELSE 188
187PRINT K,W,10*LØG10(M),R
188NEXT K
189IF A=1 THEN 190 ELSE 195
190PRINT"K,W,10LØGM,R"
195PRINT "P,R/P,10LØG(R/P)"
200PRINT R, R/P,10*LØG10(R/P)
205END
```

FIGURE A-7 LISTING OF PROGRAM INT2

```
>LOAD PFLT
>LIST
010PRINT "ENTER Z,I,L,N"
020INPUT Z,D,J,N
030PRINT "ENTER A(1) TO A(Z)"
035FOR J=1 TO Z
040INPUT A(J)
045NEXT J
050PRINT "ENTER B(1) TO B(Z)"
055FOR J=1 TO Z
060INPUT B(J)
065NEXT J
070PRINT "ENTER C(1) TO C(Z)"
075FOR J=1 TO Z
080INPUT C(J)
085NEXT J
090FOR J=1 TO Z
095N(J)=((B(J)↑2-C(J)↑2)↑2+4*(B(J)*C(J))↑2)*(A(J)↑2)
100G(J)=A(J)↑2
105E(J)=B(J)↑2-C(J)↑2
110F(J)=4*(B(J)*C(J))↑2
115NEXT J
120FOR K=1 TO N
125W=K*D
130W2=W↑2
135FOR J=1 TO Z
140D(J)=(W2+G(J))*((W2+E(J))↑2+F(J))
145M(J)=N(J)/D(J)
150NEXT J
155M=1
160FOR J= 1 TO Z
165M=M*M(J)
170NEXT J
175PRINT10*LOG10(M(1)),10*LOG10(M(2)),10*LOG10(M(3)),10*LOG10(M(4)),10*LOG10(M)
180NEXT K
185PRINT "M1,M2,M3,M4,AND M IN DB"
190END
>
```

FIGURE A-8 LISTING OF PROGRAM PFLT

```
>LØAD RESA1
>LIST
010PRINT"THIS PRØG ALLØCATES RESØURCES IDEALLY FØR A MIX"
011PRINT"ØF RATES AND CAPACITY QUØTIENTS"
020PRINT"ENTER NUMBER ØF ACCESSES"
030INPUT N
040PRINT"ENTER DESIRED SNR"
050INPUT X
060DIM Q(1:N)
070PRINT"ENTER CAPACITY QUØTIENTS"
080FØR I=1 TØ
090INPUT Q(I)
100NEXT I
110DIMR(1:N)
120PRINT"ENTER DESIRED DATA RATES"
130FØR I=1 TØ N
140INPUT R(I)
150NEXT I
160DIM D(1:N)
170FØR I=1 TØ N
180D(I)=R(I)/Q(I)
190NEXT I
200G=0
210FØR I=1 TØ N
220G=G+D(I)
230NEXT I
240DIM B(1:N)
250FØR I=1 TØ N
254F=D(I)/G
256H=Q(I)/X
260B(I)=F*H
270PRINT I,B(I),F,H
280NEXT I
290PRINT"I","RATE","%PØWER","PEAK RATE"
300T=0
310FØRI=1 TØ N
320T=T+B(I)
330NEXT I
340PRINT T
350PRINT"SATELLITE THRØUGHPUT"
360END
```

FIGURE A-9 LISTING OF PROGRAM RESA1

Appendix B

TDMA AND FDMA PERFORMANCE WITH UNEQUAL G/T USERS

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Appendix B

TDMA AND FDMA PERFORMANCE WITH UNEQUAL G/T USERS

The object of this appendix is to examine performance of TDMA and FDMA when users have different values of G/T. The performance criteria is throughput. It is assumed that each user has the same baseband data rate. The following is a list of important definitions:

Capacity quotient $Q_i = (G/T)_i \quad i = 1, n$

n = Total number of users

$$Q_{max} = \max (Q_1, Q_2, \ldots Q_n)$$

$$\Delta Q_i = \frac{Q_{max}}{Q_i} \text{. Therefore,}$$

$$\Delta Q_i \geq 1 \quad . \tag{B-1}$$

1. TDMA System

Assumption: Let R_i = RF burst rate. Then

$$R_i = kQ_i \tag{B-2}$$

where k is a constant.

Each user is allocated a time slot T_i where

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$$T_i = \frac{k^1}{R_i} \tag{B-3}$$

$$T_F = \sum_{i=1}^{n} T_i \quad . \tag{B-4}$$

Define

$$R = max\ R_i = k\ Q_{max} \quad . \tag{B-5}$$

Let TP = throughput in bits per second. Then

$$TP = \sum_{i=1}^{n} R_i \frac{T_i}{T_F} \quad . \tag{B-6}$$

From Eqs. (B-3) and (B-4),

$$\frac{T_i}{TF} = \frac{T_i}{\sum_j T_j}$$

$$= \frac{\frac{1}{R_i}}{\sum_j \frac{1}{R_j}} \quad . \tag{B-7}$$

From Eqs. (B-2), (B-1), and (B-5),

$$\frac{1}{R_i} = \frac{1}{kQ_i} = \frac{\Delta Q_i}{k\ Q_{max}} = \frac{\Delta Q_i}{R} \quad . \tag{B-8}$$

Therefore,

$$\frac{T_i}{T_F} = \frac{\Delta Q_{i/R}}{\sum_j \frac{\Delta Q_j}{R}} = \frac{\Delta Q_i}{\sum_{j=1}^{n} \Delta Q_j} \quad . \qquad \text{(B-9)}$$

From Eqs. (B-8) and (B-9),

$$TP = \sum_{i=1}^{n} R_i \frac{T_i}{T_F}$$

$$= \sum_{i=1}^{n} \frac{R}{\Delta Q_i} \frac{\Delta Q_i}{\sum_{j=1}^{n} \Delta Q_j}$$

$$= \frac{nR}{\sum_{j=1}^{n} \Delta Q_j} \quad . \qquad \text{(B-10)}$$

Throughput (TP) is maximized at $\Delta Q_j = 1$ (all j). Note that the throughput is determined by the average capacity quotient difference factor.

2. FDMA System

Assumption: All R_i equal

$$P_i \propto \frac{R_i}{Q_i}$$

or

$$R_i = kP_iQ_i \quad . \qquad \text{(B-11)}$$

We have

$$Q_i = \frac{Q_{max}}{\Delta Q_i} \quad . \tag{B-12}$$

Define:

$$R = k\ P_{max}\ Q_{max} \tag{B-13}$$

where P_{max} = down-link power if there is only one user, and

$$P_{max} = \sum_{j=1}^{n} P_i \quad . \tag{B-14}$$

From Eq. (B-13) we have

$$k = \frac{R}{P_{max}\ Q_{max}} \quad . \tag{B-15}$$

From Eqs. (B-11), (B-12), and (B-15),

$$R_i = k\ P_i\ \frac{Q_{max}}{\Delta Q_i} \tag{B-16}$$

$$= \frac{R}{P_{max}\ Q_{max}} \cdot P_i\ \frac{Q_{max}}{\Delta Q_i}$$

$$= \frac{R}{\Delta Q_i} \cdot \frac{P_i}{P_{max}} \quad . \tag{B-17}$$

From Eq. (B-16),

$$P_i = \frac{R_i \; \Delta Q_i}{k \; Q_{max}}$$

$$\sum P_i = \frac{R_i}{k \; Q_{max}} \sum \Delta Q_i \qquad ,$$

since all R_i are equal. Therefore,

$$\frac{P_i}{\sum_j P_j} = \frac{\Delta Q_i}{\sum_j \Delta Q_j}$$

$$R_i = \frac{R}{\Delta Q_i} \cdot \frac{\Delta Q_i}{\sum_j \Delta Q_j} = \frac{R}{\sum_j \Delta Q_j} \qquad .$$

Thus,

$$TP = \sum_{i=1}^{n} R_i = nR_i = \frac{nR}{\sum_{j=1}^{n} \Delta Q_j} \qquad . \qquad \text{(B-18)}$$

Consequently, FDMA throughput is also determined by the average capacity quotient difference factor.

Appendix C

EFFECT OF POWER VARIATIONS ON NETWORK-TIMING-CONTROL SYSTEM

Appendix C

EFFECT OF POWER VARIATIONS ON NETWORK-TIMING-CONTROL SYSTEM

The difficulty in establishing reliable network timing (particularly the local ranging function) without some power control can be best appreciated by the following example. Consider the problem for the lowest-sensitivity receiver where the capacity quotient may be as low as Q = 64 dB(Hz). Suppose there are 20 accesses and that it is desired to keep the sum of the low-level ranging signals 20 dB below the QPSK data signal so as not to create significant degradation to the data. This means that the effective capacity quotient for the ranging signal to the least sensitive receiver may be as low as 31 dB(Hz). Fortunately, for stability reasons the closed-loop equivalent noise bandwidth of the local ranging loop can be no greater than 1 Hz. Thus, if one were to neglect the effect of the data signals acting as interference, the SNR within this loop bandwidth would be approximately 31 dB.

Experience with delay-lock-loop tracking systems indicates that the threshold occurs at approximately 13.5 dB. To provide adequate margin and avoid frequent loss of lock, a value of approximately 20 dB is reasonable. Thus, we see that we have a margin of about 10 dB to account for the effect of interference from the data signals. Recall that the loop SNR is given by

$$SNR_r = P_r/[P_I \cdot (b/W) + N_o b] \tag{C-1}$$

where P_r is the power in the weak ranging signal, P_I is the interference power from the other ranging signals and the data signals, b is the

closed-loop equivalent noise bandwidth of the tracking system, W is the RF bandwidth, and N_o is the one-sided noise power spectral density. The previous analysis established that $P_r/N_o b \approx 1000$ and that we desire $SNR_r > 100$. Thus, it will be necessary to show that we can make

$$(P_I/P_r)\,(b/W) < 0.01 \quad . \qquad \text{(C-2)}$$

Since the one ranging signal is of such low power, essentially the entire satellite power acts as interference. Consequently, $P_I/P_r \approx 33$ dB. Thus, to achieve the desired inequality (C-2) it is necessary for the processing gain to be greater than 53 dB. Since b is approximately 1 Hz, the theoretically required bandwidth $W \geq 200$ kHz is quite feasible. The major problem is that, independent of how large we choose W, it is very difficult to achieve 53 dB in processing gain because of practical limitations.*

Consequently, one finds that the system operation is marginal. Serious fades in the desired ranging signal or increases in the data signal levels will cause the loss of lock in the weakest local ranging link. Thus, such a TDMA system is not immune to the effects of imbalance in signal powers, and power control should be considered.

---

*Practical factors such as frequency instability, phase jitter, polarity uncertainty due to data modulation, and component limitations will prevent one from achieving a processing gain inversely proportional to a bandwidth of 1 Hz. For example, the frequency-uncertainty and data-modulation effects will call for a system performance similar to that achieved by bandpass correlator systems. With such a system one might expect the processing gain to be related to a predetection bandwidth of approximately 100 Hz.

Appendix D

DIURNAL PATH-DELAY VARIATION

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DIURNAL PATH-DELAY VARIATION

With present military satellite systems the satellite orbits may be assumed to have an eccentricity of 0° and an inclination of 3°.* TELESAT, the Canadian domestic satellite system, uses station-keeping to keep the inclination angle with respect to the equatorial plane within 0.1°.[10] The peak-to-peak diurnal variation of path delay is presented in Table D-1 as a function of ground-station location for these two situations. The apogee of the satellite orbit is assumed to be on the equator at 170°E. Since the time of day may be assumed to be known relatively well, one should be able to predict the path delay to the satellite to within 10 μs for the improved satellite orbit. If a minimum time slot size of 20 μs is guaranteed for initial acquisition, then it should be possible to transmit a 10-μs sounding pulse for initial ranging without interfering with other accesses. This arrangement can thereby avoid the power-control problem associated with simultaneous data and low-level ranging signals.

By contrast, the problem with the conventional orbit is 30 times more serious. The required time-slot width (to guarantee that the initial ranging pulse falls within it) would be at least 300 μs. This value is unreasonably large for conventional frame rates, which are typically 1200 fps or higher.

---

*An eccentricity of 0.001 degrees results in a 46 n mile diurnal path delay variation. However, this path delay variation is relatively independent of terminal location. The maximum differential path delay variation (the important quantity for network timing) is 3.2 μs. Thus, the orbit eccentricity is a negligible source of timing error.

Table D-1

DIURNAL PATH-DELAY VARIATION

Location	Conventional (μs)	TELESAT (μs)
Hawaii	949	31
Camp Roberts	1350	40
Manila	614	21
Saigon	450	10
Tokyo	1440	48
Kodiak	1980	60
Okinawa	1094	37

Appendix E

COMPARISON OF UP-LINK ANTI-JAM CAPABILITY OF TDMA/SS WITH SSMA

Appendix E

COMPARISON OF UP-LINK ANTI-JAM CAPABILITY OF TDMA/SS WITH SSMA

Let P_r denote the total transponder power received at the ground terminal, P_S the signal power in each access at the transponder input (all signals equal), P_J the jammer power present at the transponder input, N the number of accesses, B the satellite and signal bandwidth, and b the video bandwidth.

Let us assume an ideal, "linear" transponder.

1. TDMA/SS Case

The useful signal power P received at the ground is given by

$$P = [P_S/(P_S + P_J)] \cdot (P_r/N) \quad . \tag{E-1}$$

The "noise" power at the receiver output is given approximately by

$$P_n = [P_J/(P_S + P_J)] \, [bP_r/B] + b\, N_o \quad . \tag{E-2}$$

Thus,

$$SNR\Big|_{out} = \frac{\dfrac{P_S}{P_S + P_J} \cdot \dfrac{P_r}{N}}{\dfrac{P_J}{P_S + P_J} \cdot \dfrac{P_r b}{B} + N_o b} \tag{E-3}$$

and for $P_J \gg P_S$,

$$SNR\Big|_{out} \approx \frac{\dfrac{P_S}{P_J}\;\dfrac{P_R}{N}}{\dfrac{b}{B}\,P_r + N_o b} \quad . \tag{E-4}$$

2. <u>SSMA Case</u>

The useful signal power P received at the ground is given by

$$P = [P_S/(NP_S + P_J)] \cdot P_r \quad . \tag{E-5}$$

The "noise" power at the receiver output is given approximately by

$$P_n = [P_J/(NP_S + P_J)]\;[bP_r/B] + bN_o \quad . \tag{E-6}$$

Thus,

$$SNR\Big|_{out} = \frac{\dfrac{P_S}{N[P_S + (P_J/N)]} \cdot P_r}{\dfrac{P_J}{NP_S + P_J} \cdot P_r \left(\dfrac{b}{B}\right) + N_o b} \tag{E-7}$$

and for $P_J \gg P_S$ (in fact $P_J \gg N\,P_S$),

$$SNR\Big|_{out} \approx \frac{\left(\dfrac{NP_S}{P_J}\right)\left(\dfrac{P_r}{N}\right)}{\left(\dfrac{b}{B}\right) P_r + N_o b} \quad . \tag{E-8}$$

Obviously Eq. (E-8) is improved over Eq. (E-4) by a factor of N. Thus, SSMA is N times better than TDMA/SS for very large J/S ratios under the above assumptions.

It is possible for $P_J \gg P_S$, but $P_J \approx NP_S$. In such a case the factor $P_J/(NP_S + P_J)$ in the denominator of Eq. (E-7) is approximately one-half. In this case the performance of SSMA is even better than N times as good as TDMA/SS.

Similarly, a hard limiter will tend to make SSMA even more than N times as desirable as TDMA/SS. SSMA is much less likely to encounter a 6-dB suppression effect from a sine-wave jammer.

If it were possible to perfectly exchange duty cycle for peak power, and pseudo-random time-slot hopping was employed to avoid selective jamming, then TDMA/SS would have the same performance against up-link jamming as SSMA.

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13. ABSTRACT

This report presents the results of research performed to investigate the application of the frequency-division multiple-access (FDMA) technique for providing a mix of user terminals of differing characteristics--such as data rates, transmitter powers, and receiver sensitivities--with simultaneous access capability to a limiting satellite repeater. A computer program (SYSCON) has been developed to model a PSK/FDMA satellite communication system and to optimize its performance in operation with a mix of user terminals, through selection of power and frequency plans. This capability is achieved through optimization of a norm defined as the weighted sum of the link error rates and representing the figure of merit or a measure of the system's communications performance with respect to both the power and the frequency of the links. The method of steepest descent is used for determining both power and frequency plans. It was found that through power and frequency control the limiting satellite repeater can be operated in the saturation region at substantially higher power levels (1-dB back-off) than is customary in practice.

To obtain the expression for the error rate at the input of the FDMA links, it was necessary to derive general analytic expressions for the limiter output signal, the intermodulation, and the noise components, when n signals are transmitted simultaneously through the satellite repeater. The expression for the bit error rate was then derived by assuming digital quadriphase modulation of the FDMA carriers and taking into consideration the presence of other FDMA carriers causing adjacent-channel interference, the intermodulation products generated in the limiter, and retransmitted satellite repeater noise, as well as receiver noise.

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	ROLE	WT	ROLE	WT	ROLE	WT
Limiters, frequency-division multiple access, intermodulation products, error rate, steepest descent, time-division multiple access, spread-spectrum multiple access, power control, frequency control						

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13. Abstract (Concluded)

A general comparison of the three major multiple-access alternatives--FDMA, TDMA, and SSMA--with respect to selected performance criteria that are particularly important for the military environment and for operation with a mix of users indicated that FDMA performs much better than past analyses had shown. With the same satellite power and RF bandwidth, FDMA was found to offer nearly as much satellite throughput as TDMA and considerably more than SSMA.

The report consists of three volumes. Volume One provides a summary of the study which includes a description of the analysis approach, documentation of the pertinent equations, numerical results, and conclusions. Detailed analysis and investigation of the problem areas, as well as derivations of analytical expressions, are contained in the appendices in Volume Two. Volume Three provides a technical assessment of the suitability of PSK/FDMA for operation with a mix of users and compares its performance with other multiple-access alternatives, in particular, with that of time-division multiple access (TDMA) and spread-spectrum multiple access (SSMA).